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Wound Rotor Induction Generators: Transients and Control

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2.1 Introduction

Wound rotor induction generators (WRIGs) are used as variable-speed generators connected to a strong or a weak power grid or as motors in the same conditions. Moreover, WRIGs may operate as stand-alone generators for variable speed.

In all these operational modes, WRIGs undergo transients. Transients may be caused by the following:

- Prime mover torque variations for generator mode
- Load machine torque variations for motor mode

- Power grid faults for generator mode
- Electric load variations in stand-alone generator mode

During transients, in general, speed and voltage, current amplitudes, power, torque, and frequency vary in time, until eventually, they stabilize to a new steady state.

Dynamic models for typical prime movers (Chapter 3, *Synchronous Generators*), such as hydraulic, wind, or steam (gas) turbines or internal combustion engines, are needed to investigate the complete transients of WRIGs.

An adequate WRIG model for transients is imperative, along with close-loop control systems to provide stability in speed, voltage, and frequency response when the active and reactive power demands are varied.

Typical static power converters capable of up to four-quadrant operation (super- and undersynchronous speed) also need to be investigated as a means for WRIG control for constant stator voltage and frequency, for limited variable speed range.

Vector or direct power control methods with and without motion sensors are described, and sample transient response results are given. Behavior during power grid faults is also explored, as, in some applications, WRIGs are not to be disconnected during faults, in order to contribute quickly to power balance in the power grid right after fault clearing. Let us now proceed to tackle the above-mentioned issues one by one.

2.2 The WRIG Phase Coordinate Model

The WRIG is provided with laminated stator and rotor cores with uniform slots in which three-phase windings are placed (Figure 2.1). Usually, the rotor winding is connected to copper slip-rings. Brushes

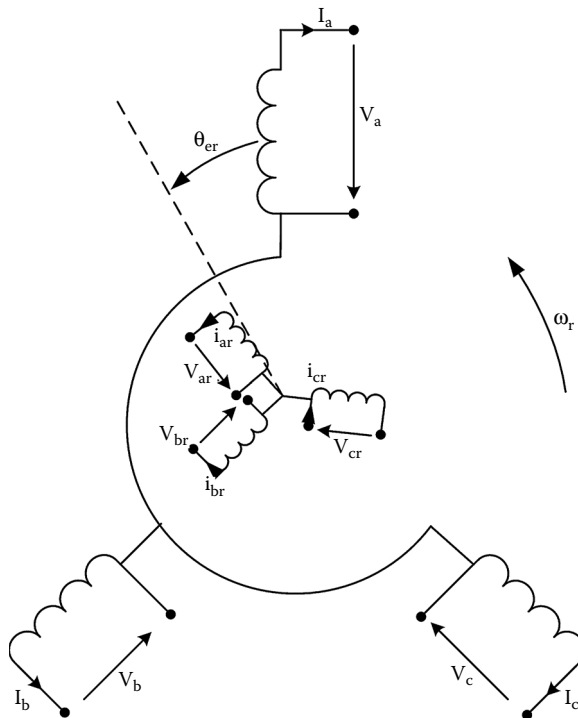


FIGURE 2.1 Wound rotor induction generator (WRIG) phase circuits.

on the stator collect (or transmit) the rotor currents from (to) the rotor-side static power converter. For the time being, the slip-ring-brush system resistances are lumped into rotor phase resistances, and the converter is replaced by an ideal voltage source.

As already pointed out in [Chapter 1](#), the windings in slots produce a quasi-sinusoidal flux density distribution in the rather uniform airgap (slot openings are neglected). Consequently, the main flux self-inductances of various stator and, respectively, rotor phases are independent of rotor position. The stator-rotor phase main flux mutual inductances, however, vary sinusoidally with rotor position θ_{cr} . The mutual inductances between stator phases are also independent of rotor position, as the airgap is basically uniform. The same is valid for mutual inductances between rotor phases. When mentioning stator and rotor phase and leakage inductances, all phase circuit parameters are included, with the exception of parameters to account for core losses (fundamental and strayload core losses). Winding strayload losses are basically caused by frequency effects in the windings and may be accounted for in the phase resistance formula. So, the phase coordinate model of WRIG is straightforward:

$$\begin{aligned} I_a R_s + V_a &= -\frac{d\Psi_a}{dt} & I_{ar} R_r + V_{ar} &= -\frac{d\Psi_{ar}}{dt} \\ I_b R_s + V_b &= -\frac{d\Psi_b}{dt} & I_{br} R_r + V_{br} &= -\frac{d\Psi_{br}}{dt} \\ I_c R_s + V_c &= -\frac{d\Psi_c}{dt} & I_{cr} R_r + V_{cr} &= -\frac{d\Psi_{cr}}{dt} \end{aligned} \quad (2.1)$$

The stator equations are written in stator coordinates, and the rotor equations are written in rotor coordinates, which explains the absence of motion-induced voltages. Generator mode association of voltage signs for both stator and rotor is evident. So, delivered electric powers are positive.

We may translate Equation 2.1 into matrix form:

$$|i_{abc,a,b,c_r}||R_{abca,b_r,c_r}| + |V_{abca,b_r,c_r}| = -\frac{d|\Psi_{abca,b_r,c_r}|}{dt} \quad (2.2)$$

$$\begin{aligned} |R_{abca,b_r,c_r}| &= \text{Diag} |R_s, R_s, R_s, R_r^r, R_r^r, R_r^r| \\ |V_{abca,b_r,c_r}| &= \text{Diag} |V_a, V_b, V_c, V_{ar}^r, V_{br}^r, V_{cr}^r|^T \\ |I_{abca,b_r,c_r}| &= \text{Diag} |I_a, I_b, I_c, I_{ar}^r, I_{br}^r, I_{cr}^r|^T \\ |\Psi_{abca,b_r,c_r}| &= \text{Diag} |\Psi_a, \Psi_b, \Psi_c, \Psi_{ar}^r, \Psi_{br}^r, \Psi_{cr}^r|^T \end{aligned} \quad (2.3)$$

The relationship between the flux linkages and currents is expressed as follows:

$$|\Psi_{abca,b_r,c_r}| = |L_{abca,b_r,c_r}(\theta_{cr})| |i_{abca,b_r,c_r}| \quad (2.4)$$

$$L_{abcq_{b_r c_r}}(\theta_{er}) = \begin{bmatrix} L_{sl} + L_{os} & \frac{L_{os}}{2} & \frac{L_{os}}{2} & M \cos \theta_{er} & M \cos(\theta_{er} + \frac{2\pi}{3}) & M \cos(\theta_{er} - \frac{2\pi}{3}) \\ -\frac{L_{os}}{2} & L_{sl} + L_{os} & \frac{L_{os}}{2} & M \cos(\theta_{er} - \frac{2\pi}{3}) & M \cos \theta_{er} & M \cos(\theta_{er} + \frac{2\pi}{3}) \\ \frac{L_{os}}{2} & \frac{L_{os}}{2} & L_{sl} + L_{os} & M \cos(\theta_{er} + \frac{2\pi}{3}) & M \cos(\theta_{er} - \frac{2\pi}{3}) & M \cos \theta_{er} \\ M \cos \theta_{er} & M \cos(\theta_{er} - \frac{2\pi}{3}) & M \cos(\theta_{er} + \frac{2\pi}{3}) & L_{rl} + L_{or} & -\frac{L_{or}}{2} & -\frac{L_{or}}{2} \\ M \cos(\theta_{er} + \frac{2\pi}{3}) & M \cos \theta_{er} & M \cos(\theta_{er} - \frac{2\pi}{3}) & -\frac{L_{or}}{2} & L_{rl} + L_{or} & -\frac{L_{or}}{2} \\ M \cos(\theta_{er} - \frac{2\pi}{3}) & M \cos(\theta_{er} + \frac{2\pi}{3}) & M \cos \theta_{er} & -\frac{L_{or}}{2} & -\frac{L_{or}}{2} & L_{rl} + L_{or} \end{bmatrix} \quad (2.5)$$

The constant mutual inductances on the stator and on the rotor are $-L_{os}/2$ and $-L_{or}/2$, because they are derived from $L_{os} \cos \frac{2\pi}{3}$ and, respectively, $L_{or} \cos \frac{2\pi}{3}$ on account of assumed sinusoidal winding (inductance) distributions. The electromagnetic torque may be derived from Equation 2.2 after multiplication by $(i_{abcq_{b_r c_r}})^T$, by using the principle of power balance:

$$\begin{aligned} |V_{abcq_{b_r c_r}}| \cdot |I_{abcq_{b_r c_r}}| &= -R_{abcq_{b_r c_r}} \cdot |I_{abcq_{b_r c_r}}|^2 - \frac{d}{dt} \left(\frac{1}{2} |I_{abcq_{b_r c_r}}|^T L_{abcq_{b_r c_r}}(\theta_{er}) I_{abcq_{b_r c_r}} \right) \\ &\quad - \frac{1}{2} |I_{abcq_{b_r c_r}}| \left| T \frac{d}{d\theta_{er}} \right| L_{abcq_{b_r c_r}} \| I_{abcq_{b_r c_r}} \| \frac{d\theta_{er}}{dt} \end{aligned} \quad (2.6)$$

The “substantial” (total) derivative d_s/dt marks the second term of Equation 2.6, which represents the stored magnetic energy variation in time, while the third term is the electromagnetic (electric) power P_{elm} , which crosses the airgap from rotor to stator or vice versa.

The first term in Equation 2.6 is the winding losses. P_{elm} should be positive for generating:

$$P_{elm} = -\frac{1}{2} |I_{abcq_{b_r c_r}}|^T \frac{d}{d\theta_{er}} \left| L_{abcq_{b_r c_r}}(\theta_{er}) \right| \| I_{abcq_{b_r c_r}} \| \omega_r = -T_e \frac{\omega_r}{p_1}; \quad \omega_r = \frac{d\theta_{er}}{dt} \quad (2.7)$$

For generating, with $P_{elm} > 0$, the electromagnetic torque T_e has to be negative (for braking the rotor):

$$T_e = + \frac{p_1}{2} \left| I_{abcq_{b_r c_r}} \right|^T \frac{dL_{abcq_{b_r c_r}}(\theta_{er})}{d\theta_{er}} \left| I_{abcq_{b_r c_r}} \right| \quad (2.8)$$

The motion equations are as follows:

$$J \frac{d\omega_r}{dt} = T_{Mech} + T_e \quad \frac{d\theta_{er}}{dt} = \omega_r \quad (2.9)$$

with $T_e < 0$ and $T_{mech} > 0$ for generating, and $T_e > 0$, $T_{mech} < 0$ for motoring. An eighth-order system of first-order differential equations was obtained. Some of its coefficients are dependent on rotor position θ_{er} that is, in time.

Such a complex model, where, in addition, magnetic saturation (implicit in L_{os} , L_{or} and M) is not easy to account for, is to be used mainly for asymmetrical (unbalanced) conditions in the power supply, in the static power converter, or in the parameters (short-circuited coils in one phase or between phases).

2.3 The Space-Phasor Model of WRIG

For stability computation or control equation system design, the phase coordinate (variable) model has to be replaced by the now widely accepted space-phasor (vector, or d - q) model obtained through the modified Park complex transformation [1]:

$$\begin{aligned}\bar{I}_s^b &= I_d^b + jI_q^b = \frac{2}{3} \left(i_a + i_b e^{j\frac{2\pi}{3}} + i_c e^{-j\frac{2\pi}{3}} \right) e^{-j\theta_b} \\ \bar{I}_r^b &= I_{dr}^b + jI_{qr}^b = \frac{2}{3} \left(i_{ar} + i_{br} e^{j\frac{2\pi}{3}} + i_{cr} e^{-j\frac{2\pi}{3}} \right) e^{-j(\theta_b - \theta_{cr})}\end{aligned}\quad (2.10)$$

The same transformation in general orthogonal coordinates, rotating at the general electric speed $\omega_b = \frac{d\theta_b}{dt}$, is valid for voltages and flux linkages $\bar{V}_s$, $\bar{V}_r$, $\bar{\Psi}_s$, $\bar{\Psi}_r$. The space phasors represent the three-phase induction machine (IM) completely, only if one more variable component in the stator and in the rotor are introduced. This is the so-called zero sequence (homopolar) component:

$$\begin{aligned}I_{os} &= \frac{1}{3}(i_a + i_b + i_c) \\ I_{or} &= \frac{1}{3}(i_{ar} + i_{br} + i_{cr})\end{aligned}\quad (2.11)$$

The zero sequence component, which is inherent to the $dq0$ model (see Chapter 6, in *Synchronous Generators*) is independent from the others and does not, in general, participate in the electromagnetic power production.

The inverse transform is as follows:

$$i_a(t) = \operatorname{Re} al \left| \bar{I}_s^b e^{j\theta_b} \right| + I_{os} \quad (2.12)$$

$$i_{ar}(t) = \operatorname{Re} al \left| \bar{I}_r^b e^{j(\theta_b - \theta_{cr})} \right| + I_{or} \quad (2.13)$$

To proceed from the phase-coordinate (variable) to the space-phasor model, let us first reduce the rotor to stator variables:

$$\begin{aligned}\frac{M}{L_{os}} &= \frac{W_2 k_{w2}}{W_1 k_{w1}} = K_{rs}; \quad \frac{L_{or}}{L_{os}} = K_{rs}^2; \quad I_{ar} = I_{ar}^r \cdot K_{rs} \\ L_{rl} &= \frac{L_{rl}^r}{K_{rs}^2}; \quad R_r = \frac{R_r^r}{K_{rs}^2}; \quad V_{ar} = \frac{V_{ar}^r}{K_{rs}}\end{aligned}\quad (2.14)$$

The transformations in Equation 2.14 “replace” the actual rotor winding with an equivalent one with the same number of turns and slots as that of the stator, while conserving the losses in the windings, the rotor electric power input, and the magnetic energy stored in the leakage inductances.

By applying the space-phasor transformations, after the reduction to stator variables, we simply obtain the voltage currents and flux/current relations for the space-phasor model:

$$\begin{aligned}
 \bar{I}_s R_s + \bar{V}_s &= -\frac{d\bar{\Psi}_s}{dt} - j\omega_b \bar{\Psi}_s \\
 \bar{I}_r R_r + \bar{V}_r &= -\frac{d\bar{\Psi}_r}{dt} - j(\omega_b - \omega_r) \bar{\Psi}_r \\
 \bar{\Psi}_s &= L_s \bar{I}_s + L_m \bar{I}_r; \quad L_s = L_{sl} + L_m; \quad L_m = \frac{3}{2} L_{os} \\
 \bar{\Psi}_r &= L_r \bar{I}_r + L_m \bar{I}_s; \quad L_r = L_{rl} + L_m
 \end{aligned} \tag{2.15}$$

The torque may be derived through the power balance principle, either from the stator or from the rotor equation:

$$\frac{3}{2} \text{Re} al \left(\bar{I}_s^* \bar{V}_s^* \right) = -\frac{3}{2} R_s I_s^2 - \text{Re} al \left(\frac{3}{2} \bar{I}_s^* \frac{d\bar{\Psi}_s}{dt} \right) - \text{Re} al \left(\frac{3}{2} j\omega_b \bar{I}_s \bar{\Psi}_s \right) \tag{2.16}$$

The electromagnetic power is represented by the last term:

$$P_{elm} = -T_e \frac{\omega_b}{p_1} = -\frac{3}{2} \omega_b \text{Im} ag \left[\bar{\Psi}_s \bar{I}_s^* \right] \tag{2.17}$$

So, the electromagnetic torque is

$$T_e = \frac{3}{2} p_1 \text{Im} ag \left[\bar{\Psi}_s \bar{I}_s^* \right] = \frac{3}{2} p_1 (\Psi_d i_q - \Psi_q i_d) = -\frac{3}{2} p_1 (\Psi_{dr} i_{qr} - \Psi_{qr} i_{dr}) \tag{2.18}$$

Again, $T_e < 0$ for generating. The factor $3/2$ stems from the complete power balance between the three-phase machine and its space-phasor model. The superscript b has been dropped for simplicity in writing.

The motion equations are the same as in Equation 2.9. The space-phasor model is to be completed with the zero sequence equations that also result from the above transformations:

$$\begin{aligned}
 I_{so} R_s + V_{so} &\approx -L_{sl} \frac{di_{so}}{dt} \\
 I_{ro} R_r + V_{ro} &\approx -L_{rl} \frac{di_{ro}}{dt}
 \end{aligned} \tag{2.19}$$

The zero sequence is irrelevant for the power transfer by the magnetomotive force (mmf) fundamental, but it produces additional stator (rotor) losses. For star connection or for symmetric transients or steady-state modes, they are, however, zero, as the sum of the phase currents is zero. The instantaneous active and reactive powers P_s , Q_s , P_r' , Q_r' , from the stator and the rotor are as follows:

$$\begin{aligned}
 P_s &= \frac{3}{2} \text{Re} al \left(\bar{V}_s \bar{I}_s^* + 2\bar{V}_{so} \bar{I}_{so}^* \right) = \frac{3}{2} \left[(V_d i_d + V_q i_q) + 2 \text{Re} al \left(\bar{V}_{so} \bar{I}_{so}^* \right) \right] \\
 Q_s &= \frac{3}{2} \text{Im} ag \left(\bar{V}_s \bar{I}_s^* + 2\bar{V}_{so} \bar{I}_{so}^* \right) = \frac{3}{2} \left[(V_d i_q - V_q i_d) + 2 \text{Im} ag \left(\bar{V}_{so} \bar{I}_{so}^* \right) \right] \\
 P_r' &= \frac{3}{2} \text{Re} al \left(\bar{V}_r \bar{I}_r^* + 2\bar{V}_{ro} \bar{I}_{ro}^* \right) = \frac{3}{2} \left[(V_{dr} i_{dr} - V_{qr} i_{qr}) + 2 \text{Re} al \left(\bar{V}_{ro} \bar{I}_{ro}^* \right) \right] \\
 Q_r' &= \frac{3}{2} \text{Re} al \left(\bar{V}_r \bar{I}_r^* + 2\bar{V}_{ro} \bar{I}_{ro}^* \right) = \frac{3}{2} \left[(V_{dr} i_{qr} - V_{qr} i_{dr}) + 2 \text{Im} ag \left(\bar{V}_{ro} \bar{I}_{ro}^* \right) \right]
 \end{aligned} \tag{2.20}$$

2.4 Space-Phasor Equivalent Circuits and Diagrams

The space-phasor equations (Equation 2.15 and Equation 2.19) may be represented in equivalent circuits that use any combination of two variables from $\bar{\Psi}_s, \bar{I}_s, \bar{\Psi}_r, \bar{I}_r$, with the other two eliminated, based on flux/current relationships (Equation 2.15). Current variables are typical:

$$\begin{aligned}\bar{\Psi}_s &= \bar{\Psi}_m + L_{sl} \bar{I}_s; & \bar{\Psi}_m &= L_m (\bar{I}_s + \bar{I}_r) \\ \bar{\Psi}_r &= \bar{\Psi}_m + L_{rl} \bar{I}_r; & \bar{I}_s + \bar{I}_r &= \bar{I}_m\end{aligned}\quad (2.21)$$

Consequently, Equation 2.15 becomes

$$(R_s + (s + j\omega_b)L_{sl})\bar{I}_s + \bar{V}_s = -L_{mt} \cdot s(\bar{I}_s + \bar{I}_r) - j\omega_b L_m (\bar{I}_s + \bar{I}_r) \quad (2.22)$$

$$(R_r + (s + j(\omega_b - \omega_r))L_{rl})\bar{I}_r + \bar{V}_r = -L_{mt} \cdot s(\bar{I}_s + \bar{I}_r) - j(\omega_b - \omega_r)L_m (\bar{I}_s + \bar{I}_r) \quad (2.23)$$

L_{mt} is the transient magnetization inductance of the WRIG. The equivalent circuit is shown in Figure 2.2.

Magnetic saturation of the main flux path is accounted for in the space-phasor model simply by the functions $L_{mt}(i_m)$ and $L_m(i_m)$, which may be determined experimentally or online. The motion-induced voltages are also visible in the coordinates system rotating at electrical speed ω_b . The coordinates system speed ω_b may be arbitrary:

- Stator coordinates: $\omega_b = 0$
- Rotor coordinates: $\omega_b = \omega_r$

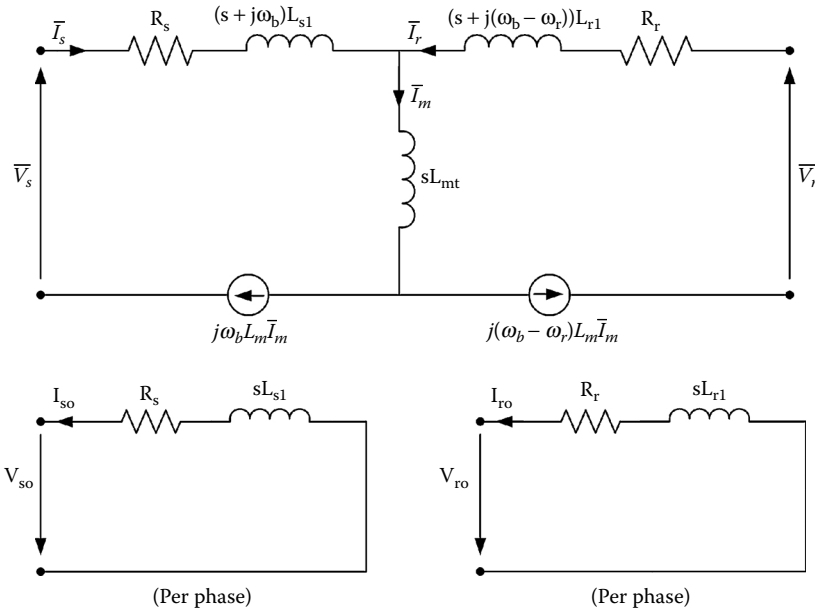


FIGURE 2.2 Space-phasor equivalent circuit of wound rotor induction generator (WRIG).

- Synchronous coordinates: $\omega_b = \omega_1$ (stator voltage or current frequency) are preferred — for steady-state and symmetrical stator voltages:

$$V_{abc}(t) = V_s \sqrt{2} \cos\left(\omega_1 t - (i-1) \frac{2\pi}{3}\right) \quad (2.24)$$

With Equation 2.10,

$$\bar{V}_s = V_s \sqrt{2} [\cos(\omega_1 - \omega_b)t - j \sin(\omega_1 - \omega_b)t] \quad (2.25)$$

Also, with

$$V_{a_r b_r c_r}(t) = V_r \sqrt{2} \cos\left((\omega_1 - \omega_r)t + \gamma - (i-1) \frac{2\pi}{3}\right) \quad (2.26)$$

$$\bar{V}_r = V_r \sqrt{2} [\cos((\omega_1 - \omega_b)t + \gamma) - j \sin((\omega_1 - \omega_b)t + \gamma)] \quad (2.27)$$

The actual frequency of the rotor voltages for steady-state ω_2 is known to be $\omega_2 = \omega_1 - \omega_r$ (see [Chapter 1](#) also).

The stator and rotor voltage space phasors have the same frequency under steady state: $\omega_1 - \omega_b$.

So, for steady state, $s =$:

- $j\omega_1$ in stator coordinates ($\omega_b = 0$)
- $j\omega_2$ ($\omega_2 = \omega_1 - \omega_r$) in rotor coordinates ($\omega_b = \omega_r$)
- 0 in synchronous coordinates ($\omega_b = \omega_1$)

At steady state, in synchronous coordinates, the WRIG voltages, currents, and flux leakages are direct current (DC) quantities. Synchronous coordinates are, thus, frequently used for WRIG control. Other equivalent circuits, with $\bar{\Psi}_s$, I_r and $\bar{\Psi}_r$, I_s pairs as variables, may also be developed but with little gain. For steady state, the space-phasor circuit has the same form irrespective of ω_b as $s = j(\omega_1 - \omega_b)$. And, if and only if magnetic saturation is ignored, $L_{mt} = L_m$ ([Figure 2.3](#)).

What differs in steady state when the reference system speed ω_b varies is the frequency of space phasors, which is $\omega_1 - \omega_b$. The equivalent circuit in [Figure 2.3](#) is similar to the per phase equivalent (phasor) circuit in [Chapter 1](#), but it has a distinct meaning. The homopolar (zero sequence) part still depends on the reference system speed ω_b . The space-phasor model may also be illustrated through the space-phasor voltage diagram ([Figure 2.4](#)).

Consider $\cos\phi = 1$ in the stator, generating operation under synchronous speed. (Active and reactive powers are absorbed through the rotor and delivered through the stator.) Consequently, the phase angle between rotor voltage and current ϕ_2 is $180^\circ < \phi_2 < 270^\circ$. It is zero between the stator voltage and current, as $\cos\phi_1 = 1$. Also, as the magnetization is produced through the rotor, $I_r > I_s$. For negative torque, $\bar{\Psi}_s$ is ahead of I_s , and $\bar{\Psi}_r$ is behind I_r for the same situation. This makes the drawing of the space-phasor diagrams during transients fairly easy, especially if we fix the coordinate system space ω_b , for example, $\omega_b = \omega_1$.

The machine is overexcited, as $\bar{\Psi}_r > \bar{\Psi}_s$, to produce unity power factor in the stator, at steady state. During steady state, for synchronous coordinates ($\omega_b = \omega_1$), d/dt terms, along the stator and rotor, flux linkage derivatives are zero.

One more way to represent the WRIG transients is with the structural (block) diagram. To derive it, we have to choose the pair of variables. The stator and rotor flux linkage space phasor $\bar{\Psi}_s$ and $\bar{\Psi}_r$ seem

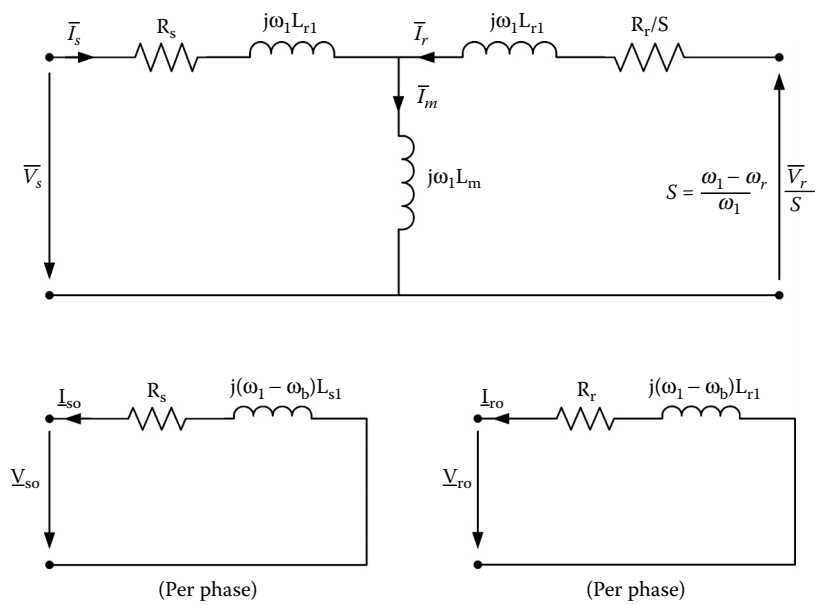


FIGURE 2.3 Steady-state space-phasor circuit model of unsaturated wound rotor induction generator (WRIG).

to be most appropriate, as they lead to a rather simplified structural diagram. First, the stator and the rotor currents are eliminated from Equation 2.15:

$$\begin{aligned} \bar{I}_s &= \frac{\left(\frac{\bar{\Psi}_s}{L_s} - \bar{\Psi}_r \frac{L_m}{L_s L_r}\right)}{\sigma}, & \sigma &= 1 - \frac{L_m^2}{L_s L_r} \\ \bar{I}_r &= \frac{\left(\frac{\bar{\Psi}_r}{L_r} - \bar{\Psi}_s \frac{L_m}{L_s L_r}\right)}{\sigma} \end{aligned} \tag{2.28}$$

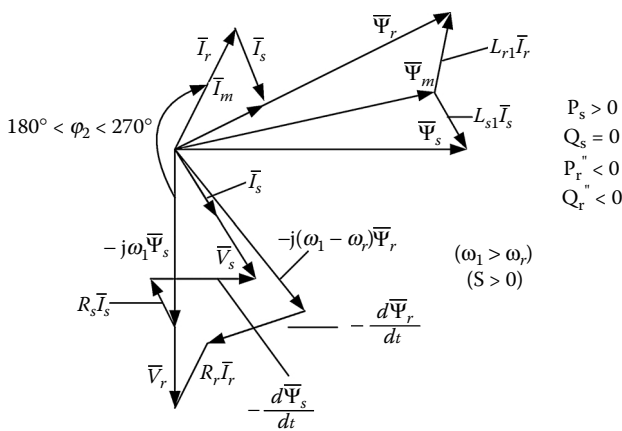


FIGURE 2.4 Space-phasor diagram of wound rotor induction generator (WRIG) for subsynchronous generator operation at unity stator power factor.

and then,

$$\tau'_s \frac{d\bar{\Psi}_s}{dt} + (1 + j\omega_b \tau'_s) \bar{\Psi}_s = -\tau'_s \bar{V}_s + K_r \bar{\Psi}_r \quad (2.29)$$

$$\tau'_r \frac{d\bar{\Psi}_r}{dt} + (1 + j(\omega_b - \omega_r) \tau'_r) \bar{\Psi}_r = -\tau'_r \bar{V}_r + K_s \bar{\Psi}_s$$

$$K_s = \frac{L_m}{L_s} \approx 0.9 - 0.97 \quad K_r = \frac{L_m}{L_r} \approx 0.91 - 0.97 \quad (2.30)$$

$$\tau'_s = \frac{L_s}{R_s} \cdot \sigma \quad \tau'_r = \frac{L_r}{R_r} \cdot \sigma \quad T_e = \frac{3}{2} p_1 \operatorname{Im} ag \left(\bar{\Psi}_s \bar{I}_s^* \right)$$

As the equations of motion are not included, Equation 2.29 represents the equation for electromagnetic transients (constant speed). Also, in general, $\omega_b = \omega_1 = \text{ct.}$ for power grid operation of a WRIG.

There is, as expected, some coupling of the stator and rotor equations through flux linkages, but the time constants involved may be called the stator and rotor transient time constants τ'_s and τ'_r , both in the order of milliseconds to a few tens of milliseconds for the entire power range of WRIGs.

As the flux linkages can vary quickly, so can the stator and rotor currents because there is a linear relationship between them (if saturation is neglected).

Equation 2.29 and Equation 2.30 lead to the structural diagram shown in Figure 2.5. The presence of the current calculator in the structural diagram is justified, because, generally, either flux-linkage or current (or torque) control is affected.

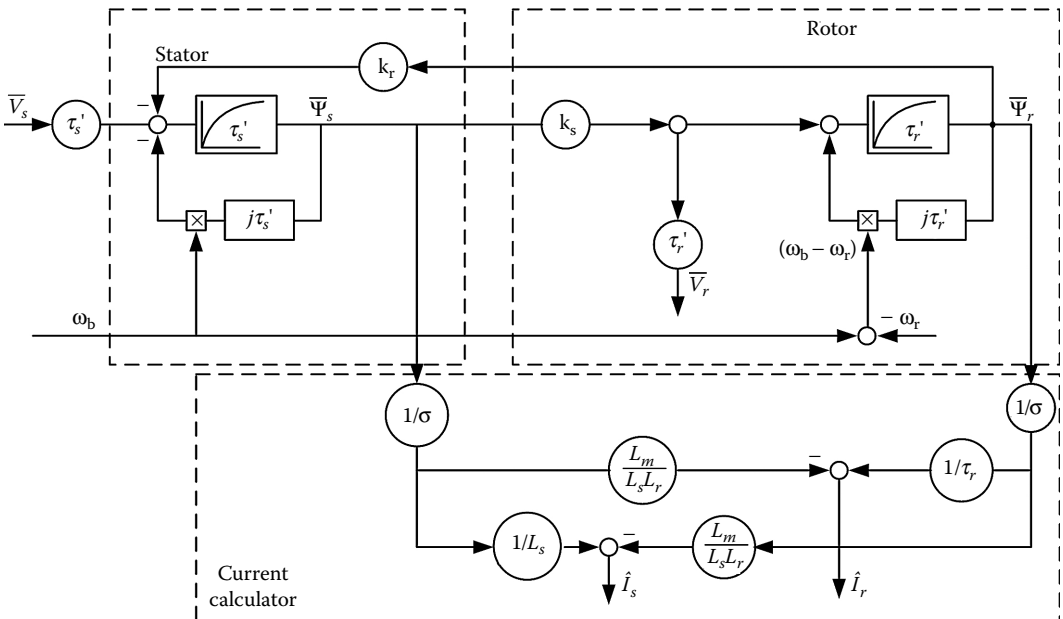


FIGURE 2.5 Wound rotor induction generator (WRIG) structural diagram.

The space-phasor (vector) model may easily be broken into d - q variables based on the following definitions (Equation 2.10):

$$\begin{aligned}
 \frac{d\Psi_d}{dt} &= -V_d - R_s i_d + \omega_b \Psi_q \\
 \frac{d\Psi_q}{dt} &= -V_q - R_s i_q - \omega_b \Psi_d \\
 \Psi_{d,q} &= L_s I_{d,q} + L_m I_{dr,q} \\
 \frac{d\Psi_{dr}}{dt} &= -V_{dr} - R_r i_d + (\omega_b - \omega_r) \Psi_{qr} \\
 \frac{d\Psi_{qr}}{dt} &= -V_{qr} - R_r i_q - (\omega_b - \omega_r) \Psi_{dr} \\
 \Psi_{dr,q} &= L_r I_{dr,q} + L_m I_{d,q} \\
 T_e &= \frac{3}{2} p_1 (\Psi_d I_q - \Psi_q I_d)
 \end{aligned} \tag{2.31}$$

$$\begin{aligned}
 \begin{vmatrix} V_d \\ V_q \\ V_{os} \end{vmatrix} &= |P(\theta_b)| \begin{vmatrix} V_a \\ V_b \\ V_c \end{vmatrix} & \begin{vmatrix} I_d \\ I_q \\ I_{os} \end{vmatrix} &= |P(\theta_b)| \begin{vmatrix} I_a \\ I_b \\ I_c \end{vmatrix} \\
 \begin{vmatrix} V_{dr} \\ V_{qr} \\ V_{or} \end{vmatrix} &= |P(\theta_b - \theta_{er})| \begin{vmatrix} V_{ar} \\ V_{br} \\ V_{cr} \end{vmatrix} & \begin{vmatrix} I_{dr} \\ I_{qr} \\ I_{or} \end{vmatrix} &= |P(\theta_b - \theta_{er})| \begin{vmatrix} I_{ar} \\ I_{br} \\ I_{cr} \end{vmatrix} \\
 P(\theta) &= \frac{2}{3} \begin{vmatrix} \cos(-\theta) & \cos\left(-\theta + \frac{2\pi}{3}\right) & \cos\left(-\theta - \frac{2\pi}{3}\right) \\ \sin(-\theta) & \sin\left(-\theta + \frac{2\pi}{3}\right) & \sin\left(-\theta - \frac{2\pi}{3}\right) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{vmatrix} \\
 |P(\theta)|^{-1} &= \frac{3}{2} |P(\theta)|^T
 \end{aligned} \tag{2.32}$$

$$\frac{d\Psi_{os}}{dt} = -V_{os} - R_s I_{os}, \quad \frac{d\Psi_{or}}{dt} = -V_{or} - R_r I_{or}; \quad \Psi_{os} \approx L_{sl} I_{os}, \quad \Psi_{or} \approx L_{sl} I_{or} \tag{2.33}$$

$$\begin{aligned}
 \frac{J}{p_1} \frac{d\omega_r}{dt} &= T_e + T_{mech}; \quad T_e < 0 \text{ for generating} \\
 \frac{d\theta_{er}}{dt} &= \omega_r; \quad \omega_r = p_1 \Omega_1 = p_1 \cdot 2\pi n
 \end{aligned} \tag{2.34}$$

The generalized matrices for stator and rotor quantities were included for completeness.

2.5 Approaches to WRIG Transients

First, we have to discriminate between the close-loop controlled and open-loop operation at constant rotor voltage or current, with slip frequency ω_2 adjusted to speed to preserve constant stator frequency $\omega_1 = \omega_2 + \omega_r$. When connected to a stiff power grid, the stator voltage is fixed in terms of amplitude and phase:

$$V_{abc} = V_s \sqrt{2} \cos \left(\omega_1 t - (i-1) \frac{2\pi}{3} \right) \quad (2.35)$$

with

$$V_{a,b,c_r} = V_r \sqrt{2} \cos \left((\omega_1 - \omega_r) t - (i-1) \frac{2\pi}{3} + \gamma_v \right) \quad (2.36)$$

for voltage “open-loop” control, and

$$I_{a,b,c_r} = I_r \sqrt{2} \cos \left((\omega_1 - \omega_r) t - (i-1) \frac{2\pi}{3} + \gamma_i \right) \quad (2.37)$$

for rotor “open-loop” control.

After using the Park transformation, in synchronous coordinates $\theta_b = \omega_1 t$, for example, for voltage “open-loop” control,

$$\begin{aligned} V_d &= V_s \sqrt{2} \\ V_q &= 0 \\ V_{dr} &= V_r \sqrt{2} \cos \gamma_v \\ V_{qr} &= -V_r \sqrt{2} \sin \gamma_v \end{aligned} \quad (2.38)$$

For constant stator voltage V_s , the influences of the phase advance of the latter with respect to stator voltage, γ_v , and the speed ω_r (or slip $S = (\omega_1 - \omega_r)/\omega_1$) on transients are paramount. In general, during transients, both electromagnetic and mechanical variables vary in time. However, in very fast transients, the speed may be considered constant; thus, the motion equations (Equation 2.34) may be ignored in the d - q model.

Laplace transform and linear control techniques may be used in such cases, but they are limited to fast rotor voltage amplitude (V_r) or phase angle (γ_r) variations. Also, a short-circuit on the stator side at WRIG terminals or somewhere along a transmission line may be approached in this way [2]. As such cases are straightforward, we leave them out here, while treating them more realistically later, in the presence of speed variation.

The linearization of the d - q model equations is used to study small deviation transients when the two motion equations are included. Subsequently, the eigenvalue method may be used to investigate small signal stability performance. Alternatively, the simplified synchronizing and damping torque coefficient method may be used for the same scope. For voltage and current (scalar) control, the eigenvalue method was successfully applied to the WRIG [3, 4] with conclusions such as the following:

- The current-controlled WRIG steady-state stability increases considerably with load, while under voltage control, only a small increase occurs.
- The power angle limits are independent of load.

- Under voltage control, for low values of rotor voltage V_r , only motor operation is stable for subsynchronous speeds ($S > 0$), and only generator operation is stable for supersynchronous speeds ($S < 0$). For current control, motor and generator operation are stable for both positive and negative slip ($S \neq 0$).

To reduce the eigenvalue computation effort, the stator resistance may be neglected. As the presence of stator resistance is known to increase damping, it follows that the results for $R_s = 0$ are conservative (on the safe side) [5].

The dynamic behavior of WRIG may also be approached by solving the d - q model equations (Equation 2.31 through Equation 2.34) through numerical methods for various perturbations [6].

The introduction of vector control, or, very recently, of direct power control of WRIGs, changed the perspectives on WRIG stability and transients, because the new controls tend to linearize the system. Very fast, almost speed-independent, active and reactive power control was obtained this way. The quality of transient response is still paramount, as there are always limitations on the mechanical side or electrical side of the WRIG system.

As the control of WRIG is strongly dependent on the static converter used on the rotor side, we will first dwell on this issue for a while.

2.6 Static Power Converters for WRIGs

The static power converter for WRIGs is connected to the rotor's three-phase windings through brushes and slip-rings. It is rated approximately at $P_{rN} = |S_{max}| P_{SN}$, where S_{max} is the maximum slip value and P_{SN} the stator rated electric power. In general, $|S_{max}| < 0.2$ to 0.3 and decreases with the power rating per unit, reaching less than 0.05 to 0.1 in the hundreds of MW power machines, to limit the static power converter rating. There are a few static power converter configurations suitable for WRIGs:

- The uncontrolled (or controlled rectifier) + current source inverter, with low cost thyristors (the DC link alternating current [AC]–AC converter)
- The cycloconverter: a voltage source commutated direct AC–AC converter with limited output/input frequency ratio $\omega_2/\omega_1 < 0.33$, with low cost thyristors
- The matrix converters: a voltage source direct AC–AC converter with insulated gate bipolar transistors (IGBTs) or integrated gate commutated thyristor (IGCTs) (or MOSFET-controlled thyristor [MCTs]) with free output/input frequency ratio
- The forced commutated rectifier-voltage source pulse-width modulator (PWM) inverter with IGBTs or IGCTs (the DC voltage link AC–AC converter)

The above-mentioned configurations differ in terms of costs, two- or four-quadrant operation under subsynchronous and supersynchronous operation, current harmonics content, and response quickness in active and reactive power control of the stator of the WRIG.

Traditionally, the first static converter introduced for WRIG (motor also) was composed of a machine-side diode rectifier with a DC choke and a source-side commutated current-source thyristor inverter (Figure 2.6).

When a diode rectifier is used on the machine side, the power flow is unidirectional — from the machine to the power grid via the step-up transformer. It means that the WRIG may operate as a motor undersynchronously ($S > 0$) and as a generator supersynchronously ($S < 0$). That is, two-quadrant operation. Also, it is not possible to go through or operate at synchronism ($S = 0$).

Finally, the current harmonics content in the rotor and in the stator is rich, and the power factor on the source side is rather modest.

For more flexibility, the current machine-side converter may be made with thyristors to allow bidirectional power flow and thus provide for four-quadrant operation: motoring and generating for both $S > 0$ and $S < 0$. The severity of harmonics content remains. Sustained operation at synchronism ($S = 0$) is not feasible, but going through synchronous speed is feasible with careful control.

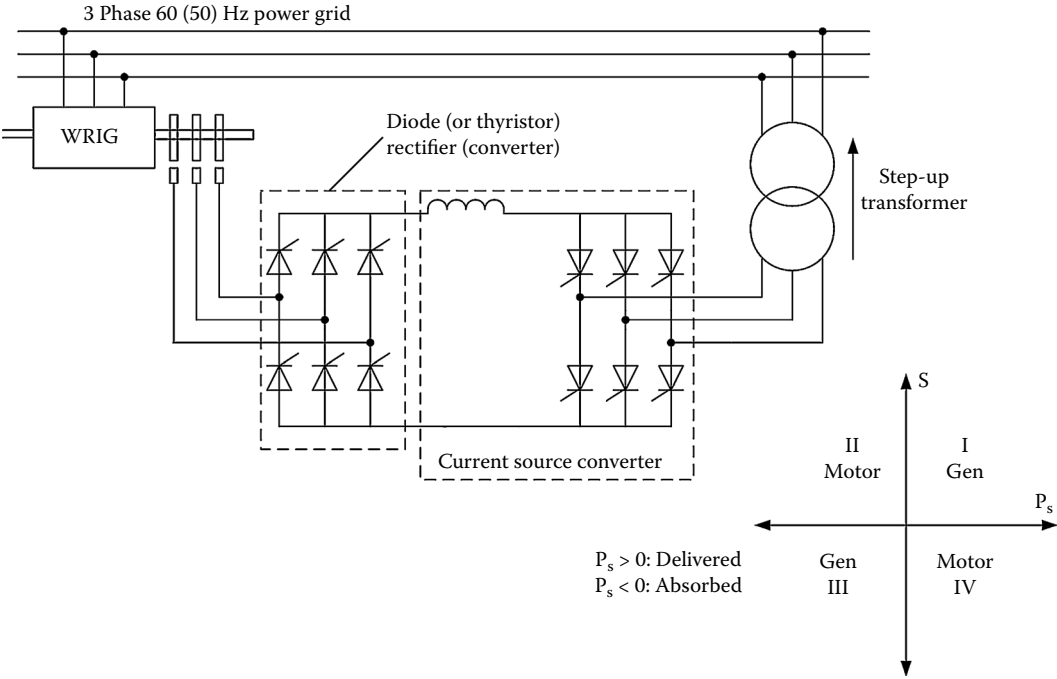


FIGURE 2.6 The direct current (DC) link alternating current (AC)–AC converter.

See References [7, 8] for a thorough analysis of the current link AC–AC converter WRIG for four-quadrant operation.

Based on Reference [7], typical experimental waveforms of thyristor voltage V_{th} , DC link voltage V_{dc} , rotor current, rotor voltage, and line current for supersynchronous and subsynchronous operation are shown in Figure 2.7.

The rotor and stator waveforms and their harmonious spectrum are shown in Figure 2.8 for $S = -0.27$.

The magnitude of supply distortion currents is less than 1% harmonics higher than fifth and seventh, but the latter may reach 10%. At low slip values, fifth and seventh rotor current harmonics may result in stator subharmonics currents. The harmonics content of supply currents increases with slip.

Overlapping over thyristor commutation and harmonics sets severe limitations on WRIG operation range. Commutation failure bars operation close to (or at) synchronous speed. The indirect AC–AC converter shown in Figure 2.6 has both component converters as source (naturally) commutated. Forced commutation may improve the situation both in terms of faster and safer commutation and in power factor, but the costs become large, and the presence of the large DC choke remains a serious drawback.

2.6.1 Direct AC–AC Converters

Direct AC–AC converters rely, again, on source commutation and are thus rather simple and inexpensive. The number of pulses may be 6, 12, or 18 to reduce current harmonics by using step-down transformers with dual or multiple delta-star (D–Y) connected secondary (Figure 2.9).

Each phase is fed from a double, controlled, rectifier — one for each current polarity — to produce operation in four quadrants (positive and negative output voltage and current). The output voltage (at rotor terminals) is “assembled” from sections of neighboring phase waveforms of the same polarity pertaining to the input (source) voltage at frequency ω_1 . The output frequency ω_2 is thus a fraction of

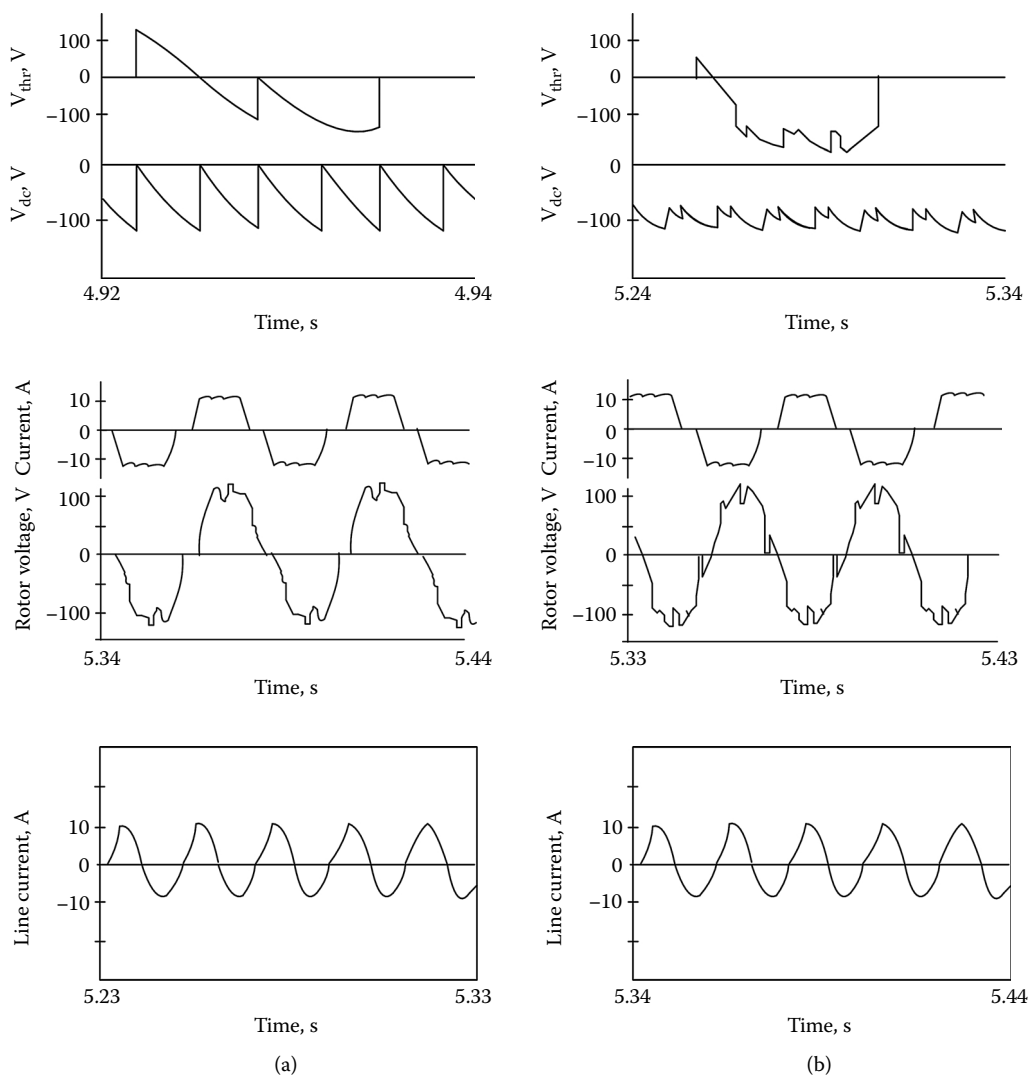


FIGURE 2.7 Direct current (DC) link alternating current (AC)-AC converter wound rotor induction generator (WRIG) typical waveforms: (a) $S < 0$ and (b) $S > 0$.

ω_1 , in practice: $|\omega_2| < \omega_1/3$. This is enough for WRIG operation, however, in case starting and synchronization as a motor are required, this is not enough, as the synchronization should take place at much higher speeds than obtainable with rotor supply and short-circuited stator motoring. Reactive power for the commutation of thyristors is drawn from the power source, and, if magnetization through rotor (unity stator power factor) is desired, overrating of the cycloconverter and transformer is required. In special configurations, ω_2 could be increased further, and this means to add capacitors for reactive power production. But in this case, the cycloconverter's essential advantages of low cost and reliability are partially lost. Harmonics with cycloconverter WRIGs are pertinently described in Reference [9]. Large power WRIGs (above 100 MW) are still provided with cycloconverters in the rotor circuit. A superior version of a direct AC-AC converter is the so-called matrix converter. The matrix converter (Figure 2.10) uses faster, bidirectional power switches and is, basically, a voltage source. The output voltage is made of sections of input voltages, intelligently “extracted” to form a PWM AC output voltage. There are some storage elements for voltage clamping (mainly) on the input side, but still, the power balance should be

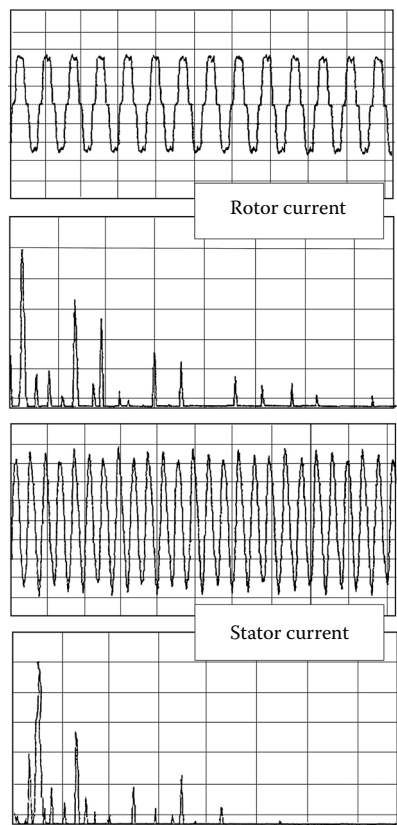


FIGURE 2.8 Wound rotor induction generator (WRIG) with direct current (DC) link alternating current (AC)–AC converter: rotor and stator current harmonics spectrum at $S = -0.27$.

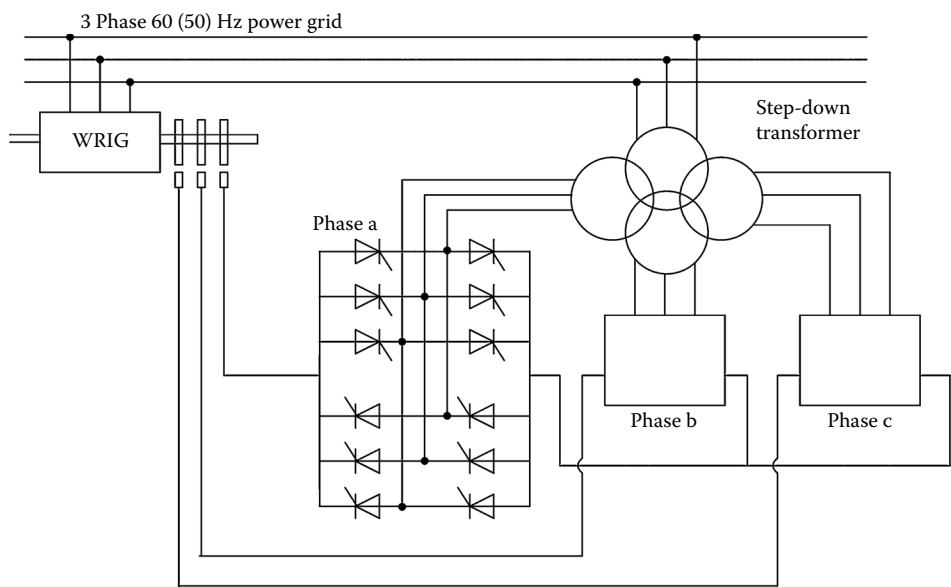


FIGURE 2.9 High-power cycloconverter for wound rotor induction generator (WRIG).

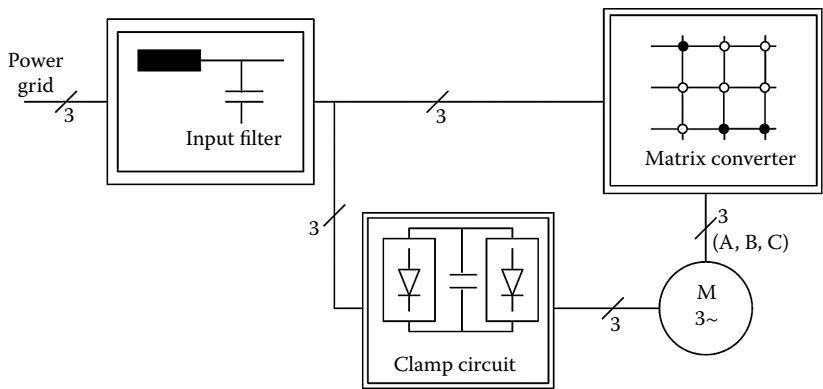


FIGURE 2.10 Typical matrix converter (with IGBTs).

provided inside each cycle in the absence of strong DC link energy storage elements. The matrix converter is claimed to show high power/volume and ultrafast power control with a controlled input power factor and low harmonics, and all this with four-quadrant natural operation and output frequency independent of ω_1 , which is essential for motors starting with WRIGs from the rotor side with a short-circuited stator.

2.6.2 DC Voltage Link AC–AC Converters

Two voltage source PWM six-leg inverters may be connected back-to-back to form a DC voltage link AC–AC indirect converter (Figure 2.11).

Bidirectional power flow is inherent, while output (rotor-side) frequency ω_2 is limited only by the switching frequency of power switches that may be gate turn-offs (GTOs), IGBTs, or IGCTs (for high power).

The two-level converter is generally used today up to 2 to 3 MW with IGBTs and line voltage outputs of 690 V. The presence of the large capacitor in the DC link provides for generous reactive power handling capacity. The rather large switching frequency (above 1 kHz) provides for lower current harmonics, both in the rotor and on the supply side.

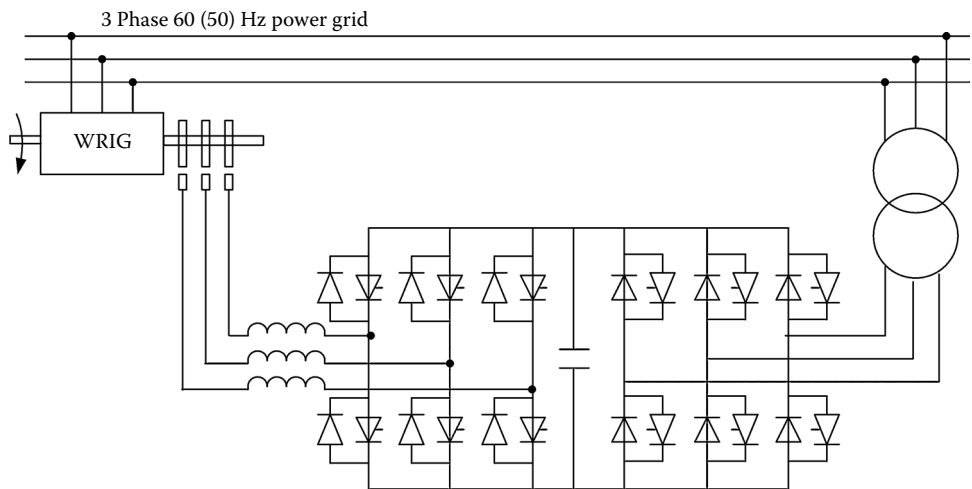


FIGURE 2.11 Two-level direct current (DC) voltage link alternating current (AC)–AC converter.

Furthermore, fast commutation of power switches provides for very fast active and reactive power response, drawing temporarily on the mechanically stored energy in the prime mover generator rotors. Consequently, intervention in the power system for power balance or voltage control is very fast, notably faster than with synchronous generators (SGs). Moreover, operation and running through synchronism ($S = 0$) comes naturally. This makes the DC voltage link AC–AC converter “ideal” for WRIGs.

But, as the power unit goes up, the maximum rotor voltage has to increase in order to keep the current through slip-rings and brushes at reasonably large levels. Voltages in the kilovolt range are typical for rotor powers up to 10 MW and above 10 kV for rotor powers above 10 MW. Multilevel DC voltage link AC–AC converters are suitable for such voltage levels. A typical three-level GTO converter is shown in Figure 2.12a and Figure 2.12b. Other “cellular-type” multilevel converters are now commercially available in the kilovolt and MW range.

To further reduce the harmonics content, it is also feasible to use two six-pulse (leg) converters in parallel on the same DC bus to work together as a 12-pulse converter when connected to a Δ, Y dual secondary step-up transformer on the source side [10].

Multiple cells in series are used to handle higher voltages, but the multilevel voltage principle holds. Such systems are also called high-voltage direct current (HVDC)-light.

After presenting the various AC–AC converters for WRIGs, it seems clear that the trend is in favor of DC voltage link AC–AC converters with IGBTs and IGCTs, up to rotor powers of MW and, respectively, tens of MW, which, in fact, cover the whole range of WRIG power per unit, up to 400 MW.

Consequently, control systems of WRIGs will be investigated in what follows only in association with DC voltage link AC–AC converters.

2.7 Vector Control of WRIG at Power Grid

Vector control stems from decoupled flux-current and torque-current control in AC drives. It resembles the principle of decoupled control of excitation and armature current in DC brush machines. Vector control is decoupled flux and torque control.

Intuitively, adding two more outer loops — one stator voltage loop to produce the reference flux current and one frequency outer loop to control generator speed — for autonomous operation is realized. When the WRIG is connected to the power grid, active and reactive powers are close-loop controlled, and they produce the reference flux and torque currents in vector control.

As motor and generator operation alternates, the control system has to be designated to handle both without hardware modifications. This makes the control system design more complicated.

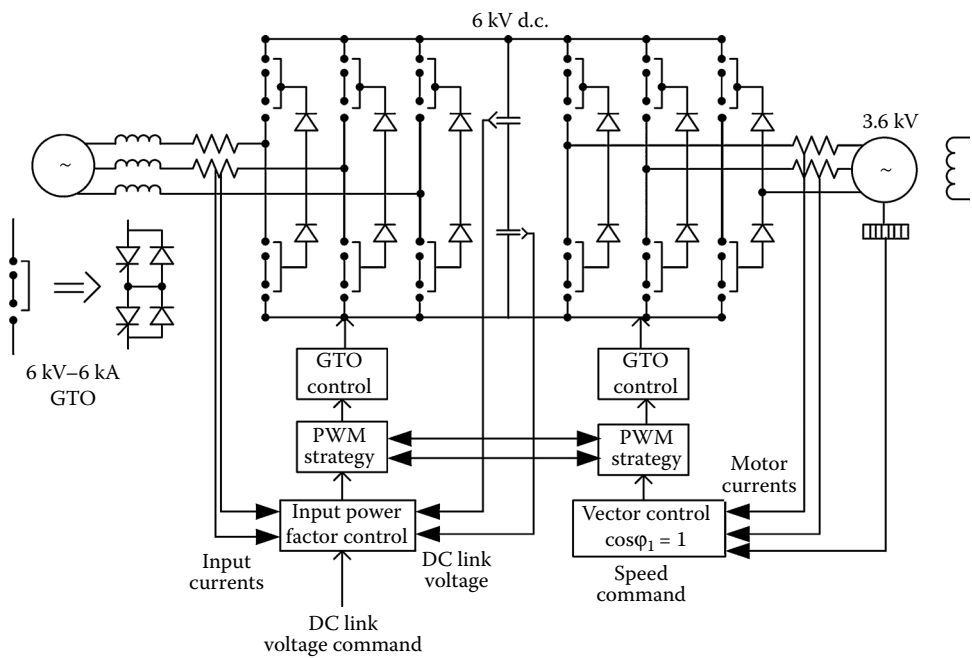
2.7.1 Principles of Vector Control of Machine (Rotor)-Side Converter

Let us restate here the WRIG space-phasor model in synchronous coordinates:

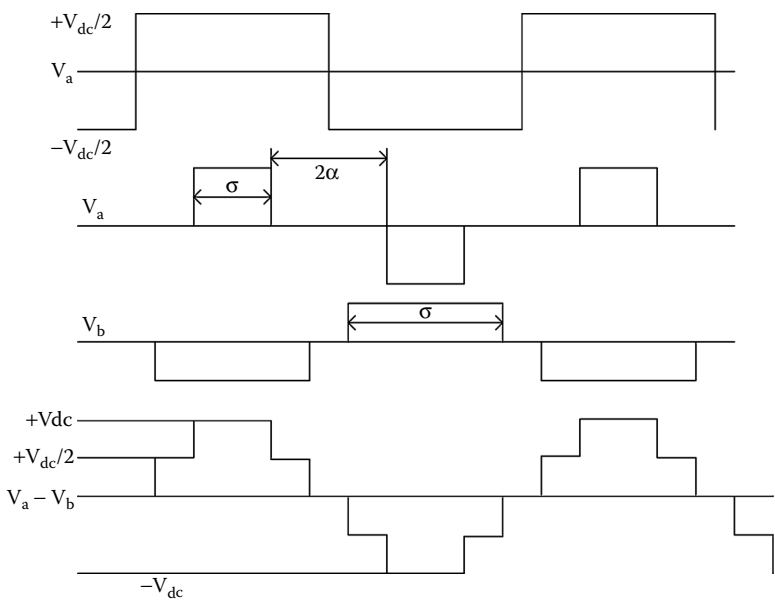
$$\begin{aligned}\bar{I}_s R_s + \bar{V}_s &= -\frac{d\bar{\Psi}_s}{dt} - j\omega \bar{\Psi}_s; & \bar{\Psi}_s &= L_s \bar{I}_s + L_m \bar{I}_r \\ \bar{I}_r R_r + \bar{V}_r &= -\frac{d\bar{\Psi}_r}{dt} - j(\omega_l - \omega_r) \bar{\Psi}_r; & \bar{\Psi}_r &= L_r \bar{I}_r + L_m \bar{I}_s\end{aligned}\quad (2.38)$$

Aligning the system of coordinates to stator flux seems most useful, as, at least for power grid operation, $\bar{\Psi}_s$ is almost constant, because the stator voltages are constant in amplitude, frequency, and phase:

$$\bar{\Psi}_s = \Psi_s = \Psi_d; \quad \Psi_q = 0; \quad \frac{d\Psi_q}{dt} = 0 \quad (2.39)$$



(a)



(b)

FIGURE 2.12 Medium-voltage direct current (DC) voltage link alternating current (AC)–AC gate turn-off (GTO) converter: (a) the configuration and (b) typical voltage waveforms.

So, the stator equation, split into d - q axes, becomes as follows:

$$\begin{aligned} I_d R_s + V_d &= -\frac{d\Psi_d}{dt} \\ I_q R_s + V_q &= -\omega_r \Psi_q \end{aligned} \quad (2.40)$$

As the stator flux Ψ_d does not vary much, $\frac{d\Psi_d}{dt} \approx 0$, and, with $R_s = 0$,

$$\begin{aligned} V_d &= 0 \\ V_q &= -\omega_r \Psi_q \end{aligned} \quad (2.41)$$

Now the active and reactive stator powers P_s , Q_s are as follows:

$$\begin{aligned} P_s &\approx \frac{3}{2}(V_d I_d + V_q I_q) = \frac{3}{2} V_q I_q = \frac{3}{2} \omega_1 \Psi_d \cdot \frac{L_m I_{qr}}{L_s}; \quad \omega_1 = \omega_r \\ Q_s &\approx \frac{3}{2}(V_d I_q - V_q I_d) = \frac{3}{2} \omega_1 \Psi_d I_d = \frac{3}{2} \omega_1 \frac{\Psi_d}{L_s} (\Psi_d - L_m I_{dr}) \end{aligned} \quad (2.42)$$

Equation 2.42 clearly shows that under stator flux orientation (vector) control, the active power delivered (or absorbed) by the stator, P_s , may be controlled through the rotor current I_{qr} , while the reactive power (at least for constant $\Psi_s = \Psi_d$) may be controlled through the rotor current I_{dr} .

Both powers depend heavily on stator flux Ψ_s and frequency ω_1 (that is on stator voltage). This constitutes the basis for vector control of P_s and Q_s by controlling the rotor currents I_{dr} and I_{qr} in synchronous coordinates.

As pulse-width modification on the machine-side converter is generally performed on rotor voltages, voltage decoupling in the rotor is required, again in synchronous coordinates.

At steady state, from Equation 2.38,

$$\bar{\Psi}_r = \frac{L_m \Psi_d}{L_s} + L_{sc} (I_{dr} + j I_{qr}); \quad \Psi_s = \Psi_d, \quad \Psi_q = 0, \quad L_{sc} = L_r - \frac{L_m^2}{L_s} \quad (2.43)$$

$$V_{dr} + j V_{qr} = -R_r (I_{dr} + j I_{qr}) - j S \omega_1 \left(\frac{L_m}{L_s} \Psi_d + L_{sc} (I_{dr} + j I_{qr}) \right) \quad (2.44)$$

From Equation 2.44,

$$\begin{aligned} V_{dr} &= -R_r I_{dr} + L_{sc} S \omega_1 I_{qr} : \quad S = 1 - \frac{\omega_r}{\omega_1} \\ V_{qr} &= -R_r I_{qr} - S \omega_1 \left(\frac{L_m}{L_s} \Psi_d + L_{sc} I_{dr} \right) \end{aligned} \quad (2.45)$$

Equation 2.45 constitutes the rotor voltage decoupling conditions. The resistance terms may be dropped as d - q rotor current closed loops are added anyway.

It also remains to estimate the stator flux linkage space-phasor Ψ_s in amplitude and instantaneous position. The inductances L_m , L_s , and L_{sc} also have to be known. The machine-side converter has to

produce the correspondents of V_{dr} and V_{qr} in rotor coordinates:

$$\begin{aligned} V_{ar}^* &= V_{dr}^* \cos(\theta_s - \theta_{er}) - V_{qr}^* \sin(\theta_s - \theta_{er}) \\ V_{br}^* &= V_{dr}^* \cos\left(\theta_s - \theta_{er} - \frac{2\pi}{3}\right) - V_{qr}^* \sin\left(\theta_s - \theta_{er} - \frac{2\pi}{3}\right) \\ V_{cr}^* &= -V_{ar}^* - V_{br}^* \end{aligned} \quad (2.46)$$

The zero voltage sequence reference voltage $V_{cr}^* = 0$. θ_s represents the stator flux angle with respect to stator phase a axis and is

$$\theta_s = \int \omega_1 dt + \theta_{s0} \quad (2.47)$$

The rotor electrical position θ_{er} is

$$\theta_{er} = p_1 \theta_r; \quad p_1 - \text{pole pairs} \quad (2.48)$$

with θ_r the rotor phase a_r axis position with respect to stator phase a axis. Under steady state, V_{dr} and V_{qr} are DC, $\theta_s = \omega_1 t$, and $\theta_{er} = \omega_1 t + \theta_{cro}$ and, as expected, the rotor voltages will have the slip frequency $\omega_2 = \omega_1 - \omega_r = S\omega_1$. The above principles of vector control are illustrated in Figure 2.13.

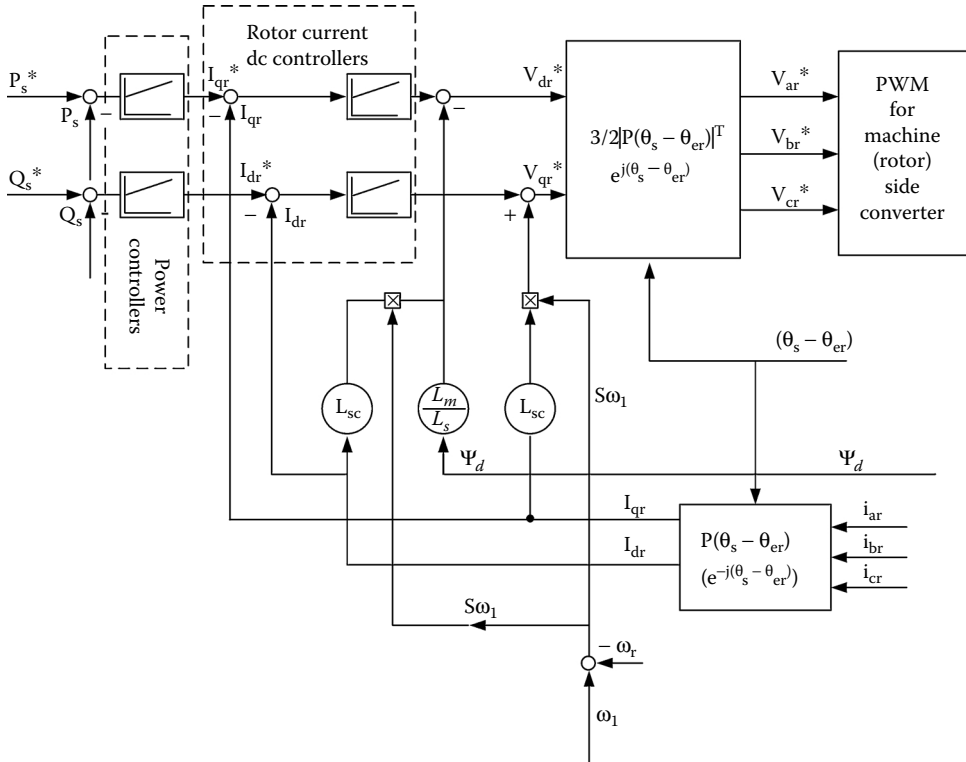


FIGURE 2.13 P_s , Q_s , vector control structural diagram for a wound rotor induction generator (WRIG).

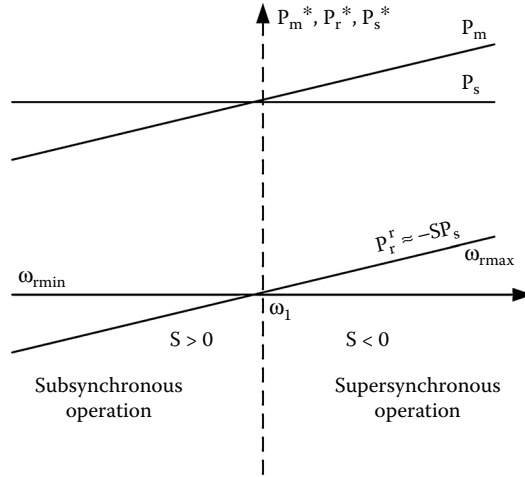


FIGURE 2.14 Power reference (envelope) division vs. speed in a wound rotor induction generator (WRIG), in generator mode.

As evident in Figure 2.13, the rotor currents have to be measured. Also, the Ψ_d and θ_s have to be estimated. The rotor position is needed. Rotor speed ω_r is also needed, both in the voltage decoupler and for prime-mover speed control. The mechanical power P_m vs. speed ω_r is obtained through an optimization criterion:

$$P_m = P_s + P_r + \sum \text{losses} \quad (2.49)$$

In general, $P_m(\omega_r)$ is linear with speed (Figure 2.14) and depends on prime-mover and WRIG characteristics.

The stator delivered power envelope increases slowly with speed, but the mechanical power varies notably as the rotor electric power changes sign at synchronous speed (ω_1).

A given total electric power requirement ($P_s + P_r \approx P_m$) asks for an optimum speed to deliver it, and the prime-mover speed generator (Chapter 3, *Synchronous Generators*) has to be able to produce it through closed-loop control. This is the second reason why speed feedback is required. Also, reference power P_s vs. speed is obtained from curves as those shown in Figure 2.14.

It is clear that either rotor electrical position or speed ω_r are measured or estimated. A resolver or a rugged (magnetic) encoder would be enough hardware to produce both θ_r and ω_r in practice. As such, a device tends to be costly and affected by noise in terms of precision. Therefore, motion estimators are preferred, leading to the so-called sensorless control. Note that as sensorless control may also be used for direct power control at the power grid or for stand-alone operation of WRIG, the estimators for sensorless control and performance with them will be treated after such control alternatives are exposed in forthcoming paragraphs.

2.7.2 Vector Control of Source-Side Converter

The source-side converter is connected to the power grid eventually via a step-up transformer in some embodiments. However, the transformer is eliminated by designing the rotor windings with a higher than unity turn ratio K_s :

$$K_{rs} \approx \frac{1}{|S_{\max}|} > 1 \quad (2.50)$$

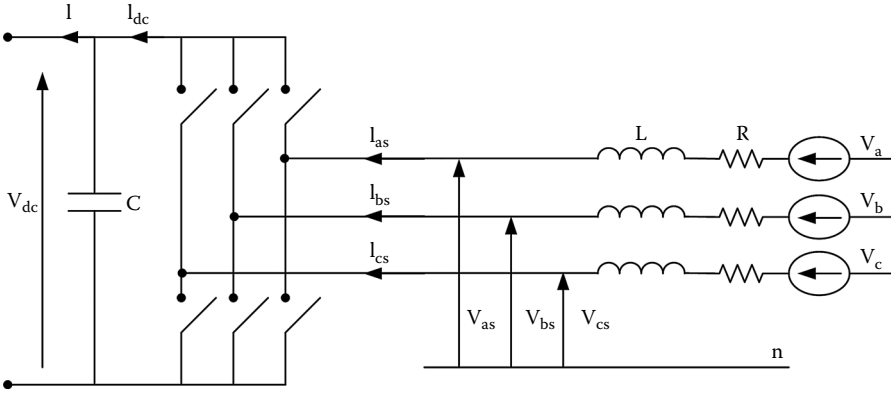


FIGURE 2.15 Source-side voltage converter.

At maximum slip, the rotor voltage equals the stator voltage. In general, the source-side voltage converter uses a power filter to reduce current harmonics flow into the power source (Figure 2.15). The presence of the source-side power filter is imperative for stand-alone operation mode.

The voltage equations across the inductors (L and R) are as follows:

$$\begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} = R \begin{bmatrix} I_{as} \\ I_{bs} \\ I_{cs} \end{bmatrix} + L \frac{d}{dt} \begin{bmatrix} I_{as} \\ I_{bs} \\ I_{cs} \end{bmatrix} + \begin{bmatrix} V_{as} \\ V_{bs} \\ V_{cs} \end{bmatrix} \quad (2.51)$$

These equations may be translated into d - q synchronous coordinates that may be aligned to axis d voltage ($V_q = 0$, $V_d = V_s$):

$$\begin{aligned} V_d &= R i_{ds} + L \frac{di_{ds}}{dt} - \omega_l L i_{qs} + V_{ds} \\ V_q &= R i_{qs} + L \frac{di_{qs}}{dt} + \omega_l L i_{ds} + V_{qs} \end{aligned} \quad (2.52)$$

where ω_l is the speed of the reference system or the supply frequency.

Neglecting the harmonics due to switching in the converter and the machine losses and converter losses, the active power balance equation is as follows:

$$V_{dc} I_{dc} = \frac{3}{2} V_d I_d = P_r; \quad V_q = 0 \quad (2.53)$$

But, with the PWM depth, m_1 , as known,

$$V_d = \frac{m_1}{2\sqrt{2}} V_{dc} \quad (2.54)$$

Then, from Equation 2.53,

$$I_{dc} = \frac{3m_1 I_d}{4\sqrt{2}} \quad (2.55)$$

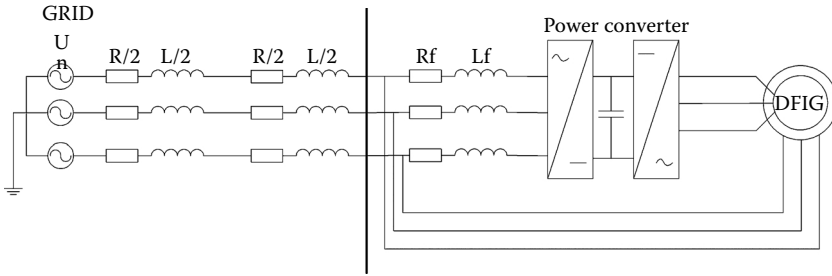


FIGURE 2.17 Wound rotor induction generator (WRIG) connected to the power grid.

Further on, the speed of ideal stator flux linkage may be calculated and filtered properly for less noise. As the stator flux Ψ_s (Ψ_q) is estimated for the vector control of the machine-side converter, this latter method may prove to be more practical. The reactive power exchange Q_r with the source-side converter is included in Figure 2.16. A complete design methodology with digital simulations and laboratory-scale results for the vector control of WRIG at power grid is given in Reference [11]. Instead of reference reactive power Q_r^* , a given small power factor angle (positive or negative) φ^* may be defined as follows:

$$I_q^* = I_d^* \tan \varphi^* \quad (2.59)$$

This solution may simplify the implementation, as only multiplication by a constant is performed. We will now present through digital simulations a description of a case study for a 2 MW wind generator application [12]. A hydraulic turbine prime mover would impose slightly different optimum mechanical power P_m vs. speed and mechanical time constants in the speed governor. It might also need motor starting for pump storage. These aspects will be treated later in this chapter.

2.7.3 Wind Power WRIG Vector Control at the Power Grid

A schematic diagram of the overall system is shown in Figure 2.17.

Two back-to-back voltage-fed PWM converters are inserted in the rotor circuit, with the supply-side PWM converter connected to the grid by a resistance inductance (RL) filter, which limits the high-frequency ripple due to switching harmonics. The complete simulation model of the system was implemented in MATLAB® and Simulink®, and its parts are explained in what follows.

2.7.3.1 The Wind Turbine Model

The wind turbine is modeled in terms of optimal tracking, to provide maximum energy capture from the wind. The implemented characteristics are presented in Figure 2.18a and Figure 2.18b.

These characteristics are based on the following:

$$P_M = \frac{1}{2} \rho_{air} \cdot C_p(\lambda, \beta) \cdot \pi \cdot R^2 \cdot U^3 \quad (2.60)$$

where

C_p is the power efficiency coefficient

U is the wind velocity

β is the pitch angle

R is the blade radius

P_M is the mechanical power produced by the wind turbine
 ρ_{air} is the air density (usually 1.225 kg/m³)

The tip speed ratio λ is

$$\lambda = \frac{\omega_r R}{U} \tag{2.61}$$

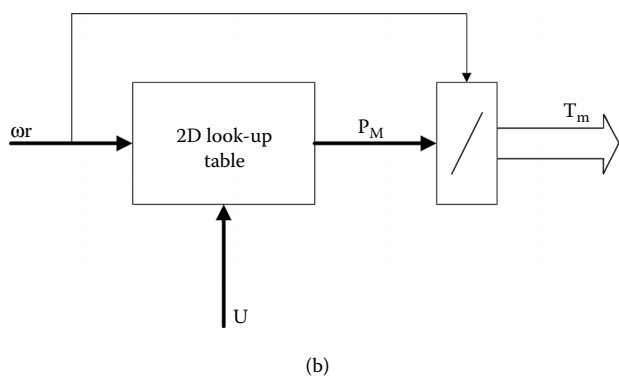
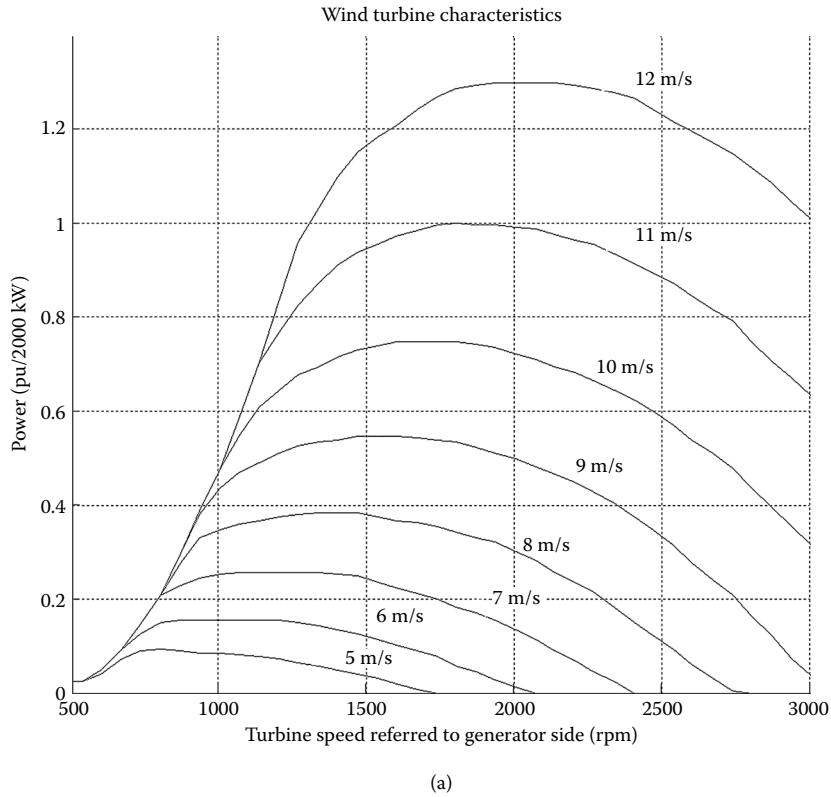


FIGURE 2.18 Implemented wind turbine characteristics: (a) aerodynamic characteristics and (b) the turbine torque model.

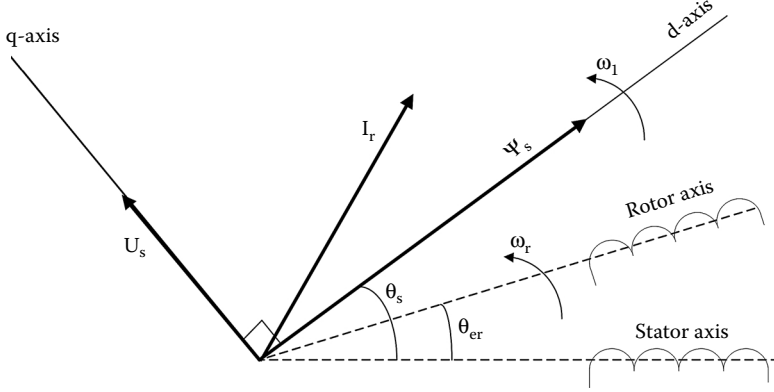


FIGURE 2.20 Location of different vectors in stationary coordinates.

with the reference frame oriented along the stator voltage. The converter is current regulated, with the direct axis current used to control the DC link voltage; meanwhile, the transverse axis current is used to regulate the displacement between the voltage and the current (and, thus, the power factor).

The angular position of the supply voltage θ_e is calculated:

$$\theta_e = \tan^{-1} \left(\frac{V_{g\beta}}{V_{g\alpha}} \right) \quad (2.64)$$

where $V_{g\alpha}$ and $V_{g\beta}$ are the stator voltages expressed in terms of α and β axes (stationary).

The design of the current controllers follows from the transfer function of the plant in the z -domain, the sampling time being 0.45 msec. For a nominal closed-loop natural frequency of 125 Hz and a damping factor of 0.8, the transfer function of the proportional integral (PI) controller is $18.69(z - 1335.4)/(z - 1)$. The current loop is designed to be much faster than the DC voltage loop and is thus considered ideal. The transfer function of the DC voltage controller results in $17.95(z - 2640)/(z - 1)$.

2.7.3.3 The Generator-Side Converter Model

The generator is controlled in a synchronously rotating d - q frame, with the d axis aligned with the stator flux vector position, which ensures a decoupled control between the electromagnetic torque and the rotor excitation contribution (Figure 2.20). In fact, decoupled active and reactive power control of the generator is obtained (Figure 2.21), as explained in a previous paragraph.

Again, knowing the transfer function of the plant and imposing the natural closed-loop frequency and the damping factor, the transfer function of the current controllers results in $12(z - 0.995)/(z - 1)$. The power loops are designed to be much slower than the current loops, and the transfer function for the power controllers are thus obtained: $0.00009(z - 0.9)/(z - 1)$.

The reference voltages generated by the current control loops with back-electromagnetic field (emf) compensation are transformed back to the rotor referenced frame and are space vector modulated to generate the voltage pulses for the inverter.

2.7.3.4 Simulation Results

Extended simulations were performed on the model thus implemented. At 1.5 sec, the wind speed is increased from 7 to 11 m/sec, then at 6 sec, the reference of active power is increased from zero to 1.2 MW, while the reference reactive power is maintained at 300 kVAR. The various stator and rotor variable waveforms are shown in Figure 2.22a through Figure 2.22f. The smooth transition through synchronous speed is noteworthy.

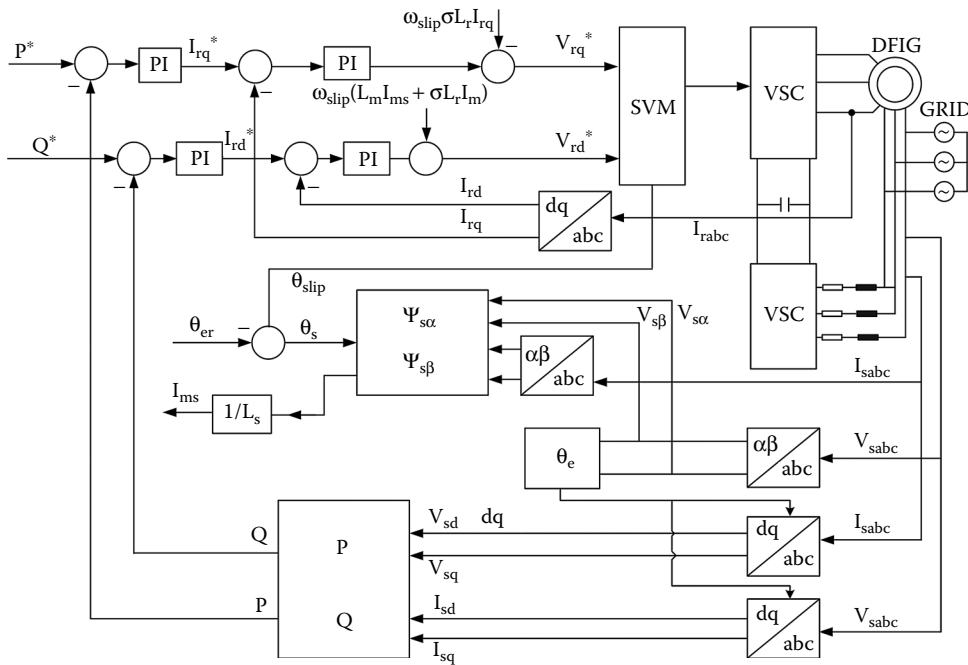


FIGURE 2.21 The block diagram of the machine-side converter control in a doubly fed wind turbine.

The stator active and reactive powers are shown in Figure 2.23a and Figure 2.23b, respectively. Decoupled control is evident as also demonstrated in Reference [13]. Safe passage through synchronous speed ($n_1 = f_1/p_1$) is visible. The system response is stable and rather quick in terms of active power control.

2.7.3.5 Three-Phase Short-Circuit on the Power Grid

In order to study the behavior of the whole system under a three-phase short-circuit on the power grid, without taking any protective measures, the grid is modeled as in Figure 2.17 and divided into two parts. The short-circuit is applied at the middle of the line. Sample results are shown in Figure 2.24a through Figure 2.24h.

It can be seen that there are two dominant current peaks, one at the initiation of the short-circuit, and one, even more severe, when the short-circuit is cleared. Two severe transients in torque also occur, and a transient in speed, which means that during the fault the electromagnetic torque is very small, and the generator accelerates. The torque of the turbine is naturally decreased during the fault, as the generator speed is increasing and the wind speed remains constant (see the characteristics in Figure 2.18).

2.7.3.6 Mechanism to Improve the Performance during Fault

To reduce the transients due to short-circuit, a few methods were investigated. The first was to reduce the active and reactive power references immediately after the fault occurs. But it did not produce any improvement.

The most efficient method, according to this study, was to limit the rotor currents. This was done by limiting the reference rotor currents, in fact, the output of the active and reactive power PI controllers at the $\pm 1.5 I_{rN}$ on the q axis and $\pm 0.5 I_{rN}$ on the d axis. The results are presented in what follows and in Figure 2.25a through Figure 2.25e.

Notable improvements can easily be noticed. First and the most important is the reduction of the second and highest rotor current peak, when short-circuit is cleared, by 50%. The stator current second

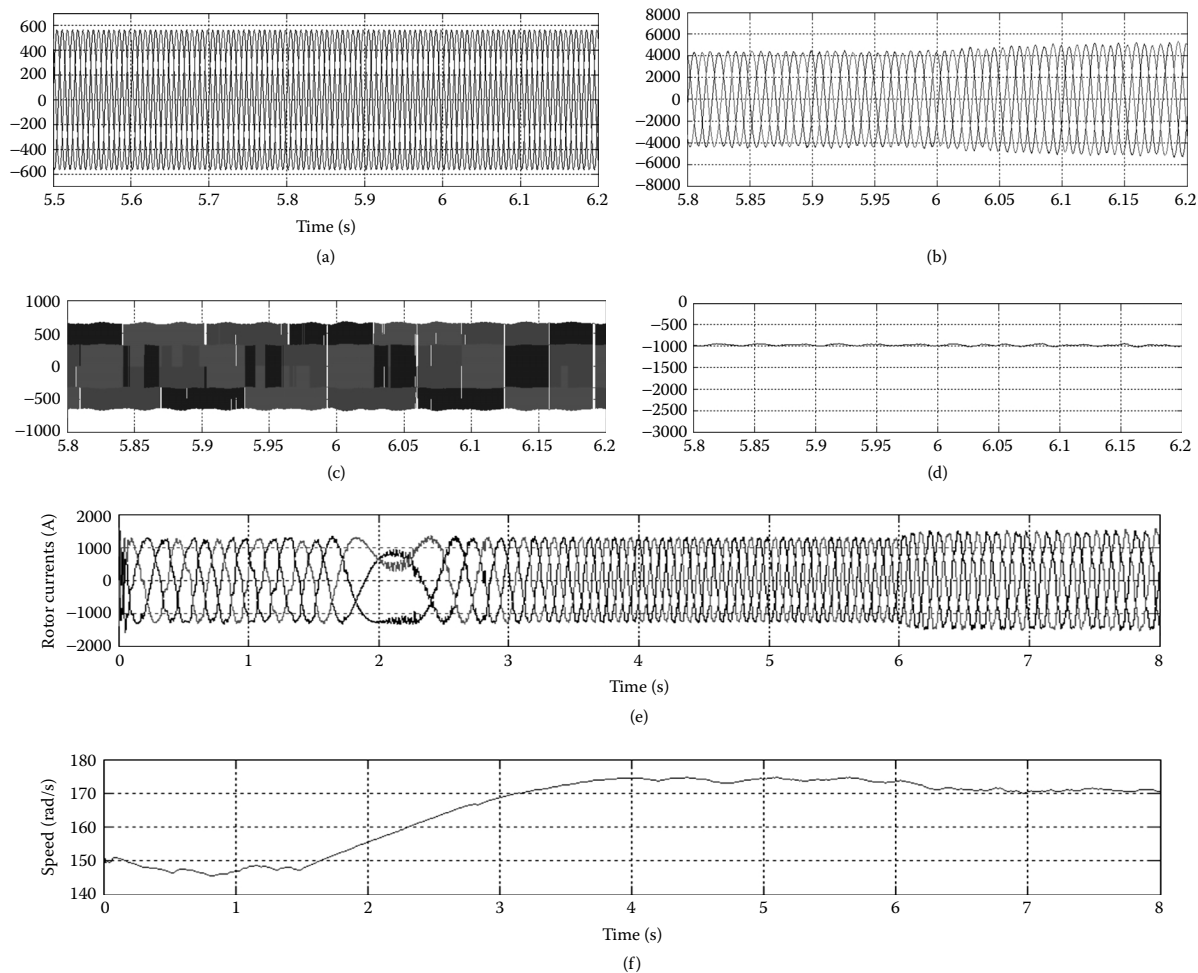


FIGURE 2.22 The generator waveforms: (a) stator voltages, (b) stator currents, (c) rotor voltages, (d) DC link voltage, (e) rotor currents, and (f) speed.

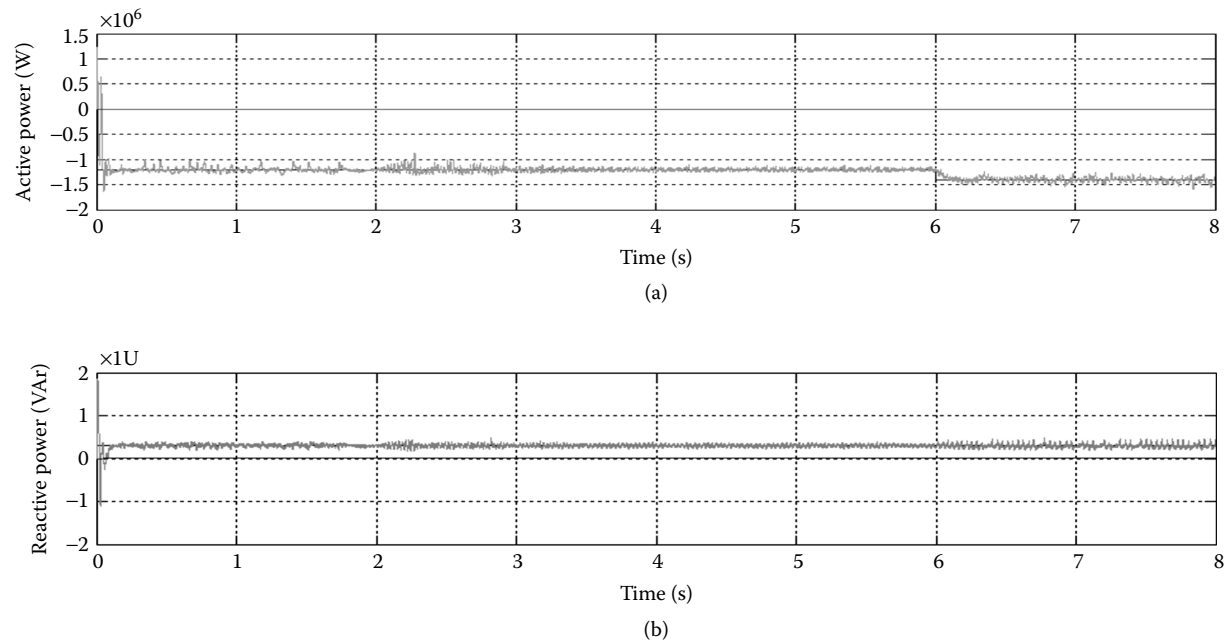


FIGURE 2.23 The (a) active and (b) reactive stator power control.

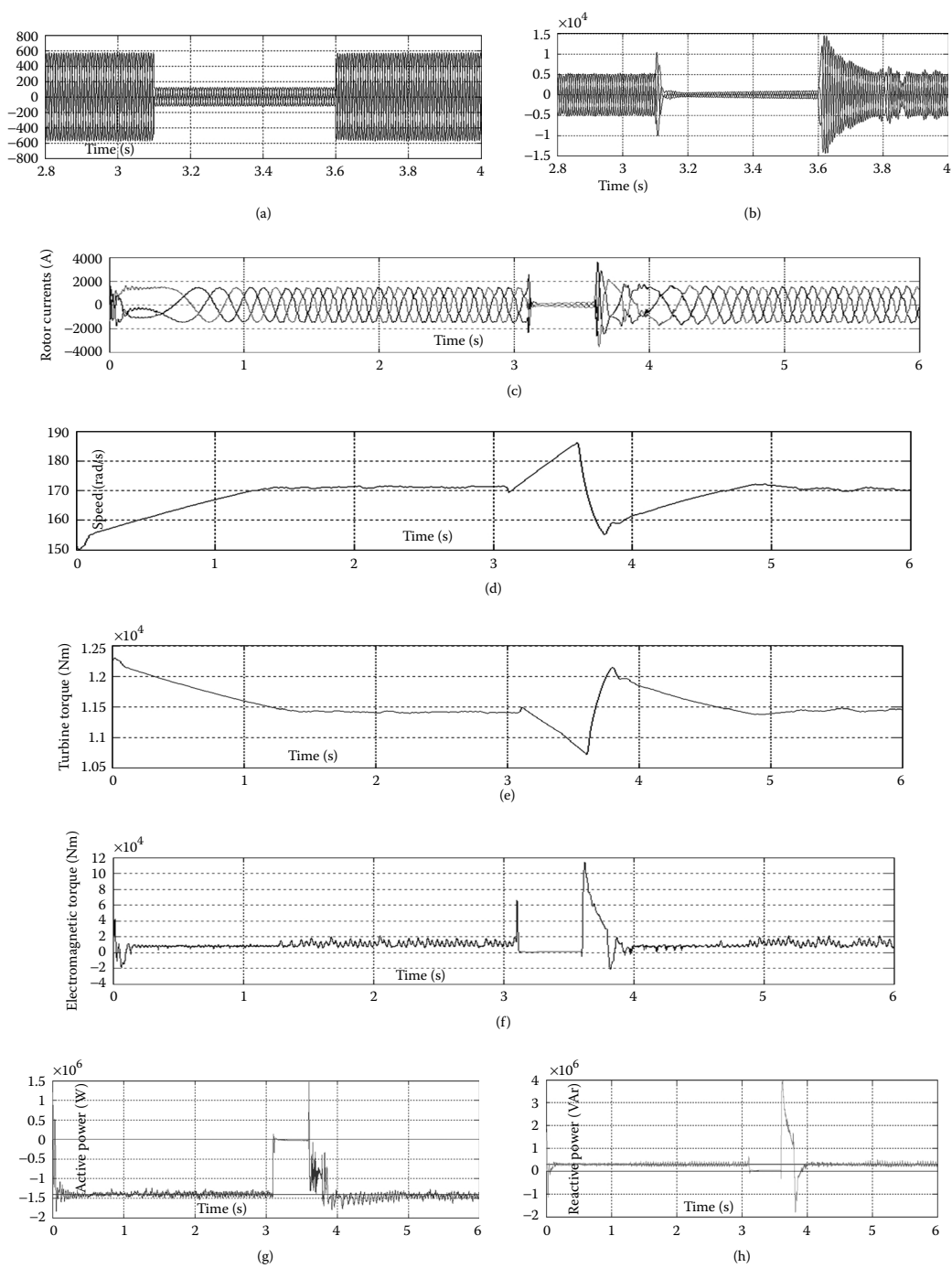


FIGURE 2.24 Three-phase short-circuit on the power grid: (a) stator voltage, (b) stator currents, (c) rotor currents, (d) speed, (e) turbine torque, (f) electromagnetic torque, (g) active power, and (h) reactive power.

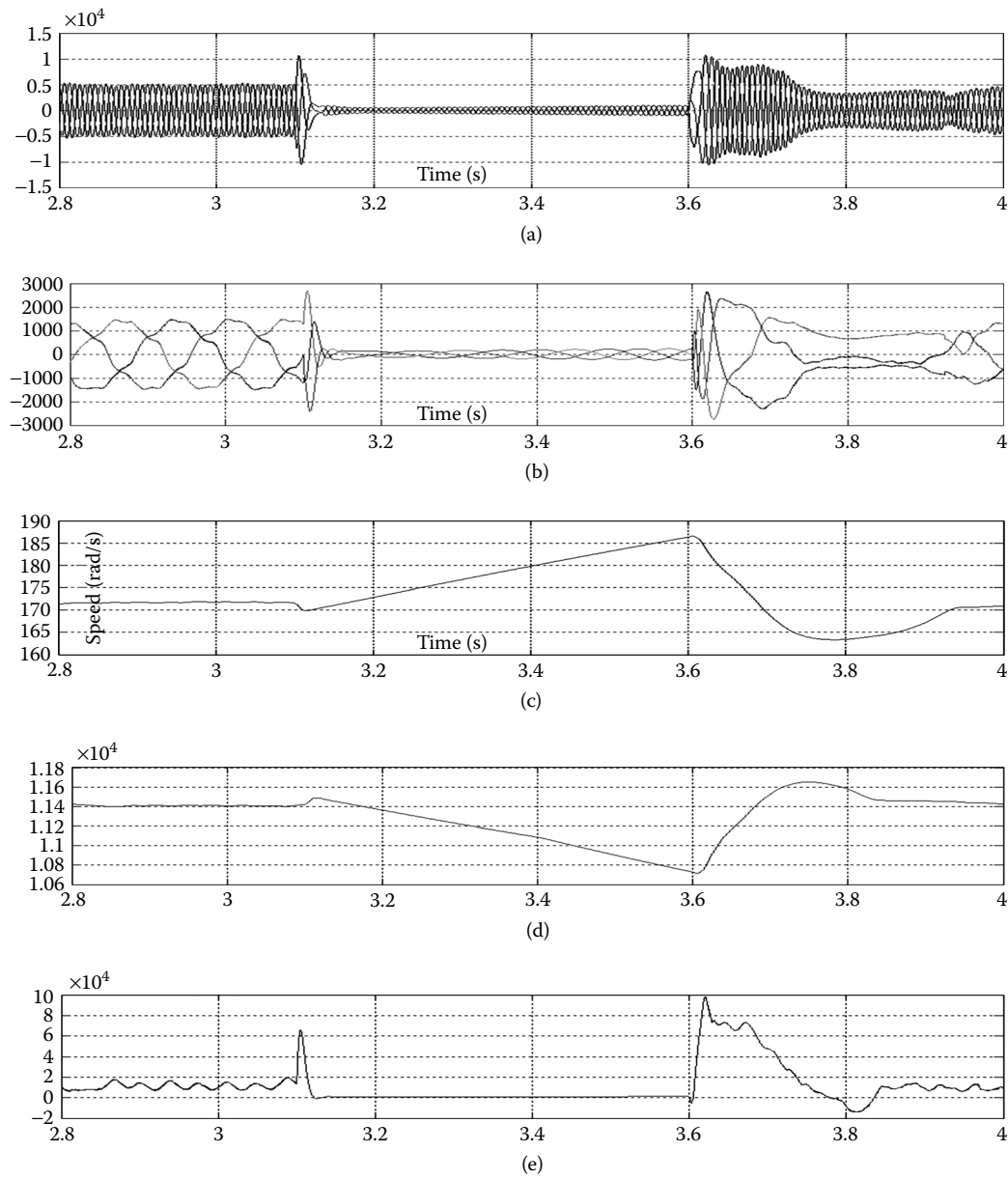


FIGURE 2.25 Three-phase short-circuit on the power grid, with limited rotor currents in the generator: (a) stator currents, (b) rotor currents, (c) speed, (d) turbine torque (Nm), and (e) electromagnetic torque.

peak is also reduced with the same amount. The acceleration of the generator is not avoided, but when the system is restoring, the deceleration is less pronounced, and the speed generator stabilizes earlier. Practically, the system is recovering 1 sec more quickly than in the case with no limitations in the rotor currents.

Note that for large power wind farms, continuous operation during short-circuit faults is required. This mode becomes more important, as the wind turbine should be able to contribute with power just after a short-circuit. The good dynamic performance and accuracy of the active and reactive power control were demonstrated. Also, the waveforms of the generator in a three-phase short-circuit are

presented. For better continuous operation during short-circuit faults, different methods were investigated. The reduction of the power references during the fault in order to prevent the rotor overcurrents and generator overspeeding was performed, but it did not produce expected results. The best situation was obtained with limitation of the rotor current references. Other methods might produce even better results.

2.8 Direct Power Control (DPC) of WRIG at Power Grid

Direct active and reactive power control of WRIGs stems from direct torque and flux control (DTFC) in AC electric drives [14]. In this perspective, vector control of active and reactive power of WRIG as presented in previous paragraphs should be considered as indirect control.

The active and reactive stator powers are calculated from measured voltages and currents and are closed-loop controlled by hysteresis or more advanced regulators. These regulators directly trigger a sequence of voltage vectors in the machine-side converter, based on power errors and also on the position of the stator flux in one of the six 60°-wide sectors. Consequently, coordinate transformation is apparently eliminated, and so is the necessity to acquire (or estimate) the rotor position.

The speed estimation (or measurement) is still needed for prime-mover control as optimum P_m (and P_s) depend on speed, and, in fact, it increases almost linearly with speed.

2.8.1 The Concept of DPC

As the stator voltage V_s is considered rather constant, in general, the stator flux Ψ_s (for constant frequency ω_1) is

$$\Psi_s \approx \frac{V_s}{\omega_1} \quad (2.65)$$

with $R_s \approx 0$. At steady state, as demonstrated for vector control, in stator flux coordinates (Equation 2.42), the active stator power P_s is dependent on I_{qp} , and reactive power Q_s is dependent on I_{dr} . The rotor currents were written in stator flux coordinates. Here we will use an alternative approach. The DPC implies directly triggering the rotor voltage vectors in rotor coordinates. Consequently, the rotor space-phasor equation is written in rotor coordinates as follows:

$$\bar{I}_r^r R_r + \bar{V}_r^r = -\frac{d\bar{\Psi}_r^r}{dt} \quad (2.66)$$

Though neglecting the resistive voltage is producing a significant error in Equation 2.66 at small values of slip S when V_r (its fundamental) is small, for the time being, we neglect it to see the greater picture of rotor flux variation that takes place along the applied rotor voltage vector:

$$\bar{\Psi}_{rf} - \bar{\Psi}_{ri} = -\int \bar{V}_r^r dt \quad (2.67)$$

The rotor flux change (increment) falls opposite to the applied voltage vector's direction, as the generator association of signs was adopted. But, the rotor flux Ψ_r is as follows:

$$\bar{\Psi}_r = \frac{L_m}{L_s} \bar{\Psi}_s + L_{sc} \bar{I}_r \quad (2.68)$$

The active and reactive stator powers P_s , Q_s (with zero stator losses) are

$$\begin{aligned} P_s &\approx -\frac{3}{2} p_1 \omega_1 \text{Imag}(\bar{\Psi}_r \bar{I}_r^*) \\ Q_s &\approx -\frac{3}{2} p_1 \omega_1 \text{Real}(\bar{\Psi}_r \bar{I}_r^*) \end{aligned} \quad (2.69)$$

with Equation 2.68, and introducing a flux power angle δ_ψ between $\bar{\Psi}_s$ and $\bar{\Psi}_r$, P_s and Q_s become

$$\begin{aligned} P_s &\approx \frac{3}{2} p_1 \omega_1 \frac{L_m \Psi_s \Psi_r}{L_s L_{sc}} \sin \delta_\psi \\ Q_s &\approx \frac{3}{2} p_1 \omega_1 \frac{L_m}{L_s L_{sc}} \Psi_r (\Psi_r - \Psi_s \cos \delta_\psi) \end{aligned} \quad (2.70)$$

The flux angle δ_ψ is positive for $\bar{\Psi}_r$ ahead of $\bar{\Psi}_s$ (Figure 2.4). Positive power means delivered power. Equations 2.69 lead us to make the following remarks:

- To increase active stator and reactive power delivery, for constant stator flux (constant V_s/ω_1), the rotor flux $\bar{\Psi}_r$ should increase in amplitude.
- To increase only active power P_s delivery, the rotor flux (power) angle δ_ψ has to be increased. For a given flux power angle δ_ψ , Equations 2.70 suggest that the reactive power generated by the machine is increased by increasing the rotor flux amplitude Ψ_r .

In other words, the rotor flux space phasor has to be decreased or increased in amplitude and accelerated or decelerated. The above rationale presupposes the same coordinates for rotor and stator flux (Figure 2.26).

The only problem is that the applied rotor voltage space phasor (vector) by the machine-side converter is in rotor coordinates. Under synchronous operation, the rotor voltage vector travels in the direction of motion (in rotor coordinates). So, accelerating the rotor flux really means accelerating it

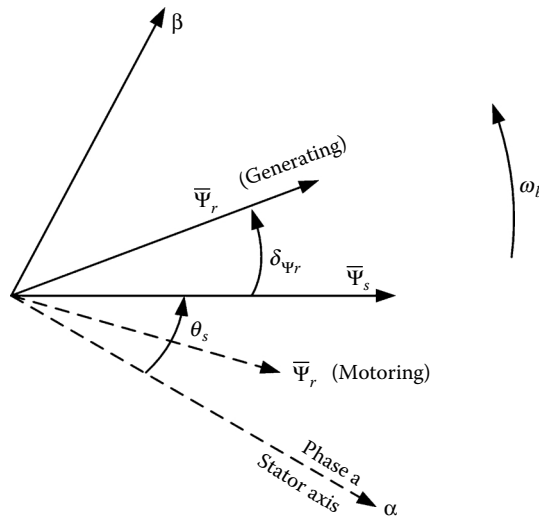


FIGURE 2.26 Stator/rotor flux angle.

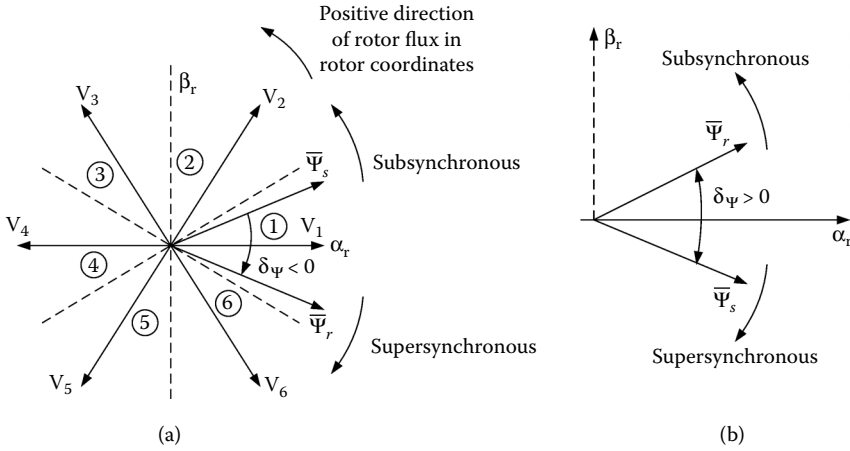


FIGURE 2.27 Stator/rotor flux vector ω_2 rotor coordinates: (a) for motoring and (b) for generating.

in the direction of motion under synchronous speed. Above synchronous speed acceleration of rotor flux has to be performed opposite to motion direction. This rationale is illustrated in Figure 2.27a and Figure 2.27b.

The rotor flux $\bar{\Psi}_r$ is estimated directly in stator coordinates from the current model:

$$\bar{\Psi}_r^s = \frac{L_r}{L_m} \bar{\Psi}_s^s - L_{sc} \bar{I}_s^s \quad (2.71)$$

with

$$\bar{\Psi}_s = -\int (\bar{V}_s - R_s \bar{I}_s) dt$$

That is, with measured stator quantities and known parameters — L_{sc} (short-circuit inductance) and coupling ratio $L_r/L_m = 1.02$ to 1.10 — the rotor flux may be estimated. Based on the sign of active and reactive power errors,

$$\begin{aligned} \varepsilon_{PS} &= P_s^* - P_s \\ \varepsilon_{QS} &= Q_s^* - Q_s \end{aligned} \quad (2.72)$$

with rotor flux vector in sector K , the application of voltage vectors $V(K+1)$ and $V(K+2)$ would increase the delivered stator active power, while the vectors $V(K-1)$ and $V(K-2)$ would reduce it. Moreover, the application of $V(k)$, $V(K-1)$, and $V(K-2)$ would decrease the delivered reactive power, while $V(k)$, $V_{(K+1)}$, $V_{(K+2)}$, and $V_{(K+3)}$ would increase it. The reactive power control is the same in motor and generator operation modes.

Notice that the order of voltage vectors is opposite in subsynchronous operation with respect to supersynchronous operation (Figure 2.27). The zero voltage vectors of the converter $V_{(0)}$, $V_{(7)}$ stall the rotor flux vector without affecting its magnitude. Their effect on active power is thus opposite in subsynchronous and supersynchronous operation. Seen from the rotor side, in subsynchronous operation, V_0 increases δ_ψ as $\bar{\Psi}_s$ travels at slip speed. Hence, the active power increases in this case.

Also, there is some influence of the zero voltage vector on the reactive power through the flux power angle δ_ψ (Equation 2.70) increment. When δ_ψ increases, $\cos \delta_\psi$ decreases, and thus, Q_s increases.

The reverse is true when δ_ψ decreases. So, the zero voltage vector effects are different in the four modes of operation (quadrants) (Table 2.1).

It becomes rather simple to adopt the adequate rotor flux sectors.

The vector flux sector may be found by the direction of change in Q_s when a certain voltage vector is applied in subsynchronous or supersynchronous modes, avoiding the use of the observer (Equation 2.70 and Equation 2.71) [15]. Table 2.2 illustrates the selection of voltage vector by ϵ_{P_s} and ϵ_{Q_s} (Equation 2.72) and the sector. Zero voltage vectors are illustrated in Table 2.3.

The influence of a voltage vector on a Q_s change sign, for a given sector, is summarized in Table 2.4.

In practice, when applying a given (known) voltage vector, the measured reactive power change sign is obtained. If the actual sign of change is opposite to that in Table 2.4, the decision is to change the sector according to Table 2.5. The source-side converter may be vector controlled, as in the previous paragraphs, for vector control. Typical experimental results are shown in Figure 2.28a and Figure 2.28b [15].

Figure 2.29a and Figure 2.29b show the vector information (the step-like function), P_s and Q_s for steady state, the passing-through synchronous speed (the rotor current), and Q_s and sector information during starting on the fly [15].

TABLE 2.1 Zero Voltage Vector Effects on P_s , Q_s

Speed	Motoring	Generating
Subsynchronous	$\delta\Psi_r \uparrow \rightarrow Q_s \downarrow$	$\delta\Psi_r \downarrow \rightarrow Q_s \uparrow$
Supersynchronous	$\delta\Psi_r \downarrow \rightarrow Q_s \uparrow$	$\delta\Psi_r \uparrow \rightarrow Q_s \downarrow$

TABLE 2.2 Active Voltage Vectors

P_s	Q_s	Sector 1	Sector 2	Sector 3	Sector 4	Sector 5	Sector 6
$\epsilon_{P_s} \leq 0$	$\epsilon_{Q_s} < 0$	V_3	V_4	V_5	V_6	V_1	V_2
$\epsilon_{P_s} \leq 0$	$\epsilon_{Q_s} \geq 0$	V_2	V_3	V_4	V_5	V_6	V_1
$\epsilon_{P_s} > 0$	$\epsilon_{Q_s} < 0$	V_5	V_6	V_1	V_2	V_3	V_4
$\epsilon_{P_s} > 0$	$\epsilon_{Q_s} \geq 0$	V_6	V_1	V_2	V_3	V_4	V_5

TABLE 2.3 Zero Voltage Vectors

Speed	Motor	Generator
Subsynchronous	$\epsilon_{Q_s} \leq 0$ and $\epsilon_{P_s} \leq 0$	$\epsilon_{Q_s} > 0$ and $\epsilon_{P_s} \leq 0$
Supersynchronous	$\epsilon_{Q_s} > 0$ and $\epsilon_{P_s} < 0$	$\epsilon_{Q_s} \geq 0$ and $\epsilon_{P_s} < 0$

TABLE 2.4 Q_s Changes Sign (Expected)

Sector	V_0	V_1	V_2	V_3	V_4	V_5	V_6	V_7
1	0	+	+	−	−	−	+	0
2	0	+	+	+	−	−	−	0
3	0	−	+	+	+	−	−	0
4	0	−	−	+	+	+	−	0
5	0	−	−	−	+	+	+	0
6	0	+	−	−	−	+	+	0

Source: Adapted from R. Datta, and V.T. Ranganathan, *IEEE Trans.*, PE-16, 3, 2001, pp. 390–399.

TABLE 2.5 Direction of Sector Change (Advance)

Sector	V_0	V_1	V_2	V_3	V_4	V_5	V_6	V_7
1	0	0	-1	+1	0	-1	+1	0
2	0	+1	0	-1	+1	0	-1	0
3	0	-1	+1	0	-1	+1	0	0
4	0	0	-1	+1	0	-1	+1	0
5	0	+1	0	-1	+1	0	-1	0
6	0	-1	+1	0	-1	+1	0	0

Source: Adapted from R. Datta, and V.T. Ranganathan, *IEEE Trans.*, PE-16, 3, 2001, pp. 390–399.

During starting on the fly, at the beginning, the sector information is not correct (Figure 2.29c), but it comes “to its senses” within 1 msec. A safe run through synchronous speed is visible.

More advanced P_s and Q_s controllers may be used, with a space vector modulator, to improve rotor current waveforms and make the DPC strategy even better, while still keeping it simple enough and robust.

Note that apart from vector control and DPC, feedback linearized control was recently proposed for the machine-side converter with promising results [16].

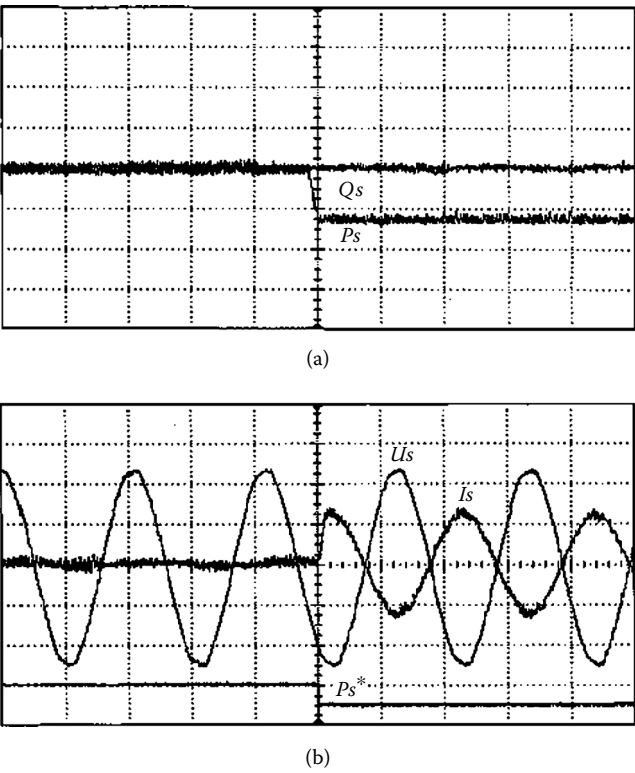


FIGURE 2.28 Direct power control (DPC) of wound rotor induction generator (WRIG): (a) P_s step variation from 0 to -0.5 P.U., $Q_s = 0$ and (b) V_s , I_s transients.

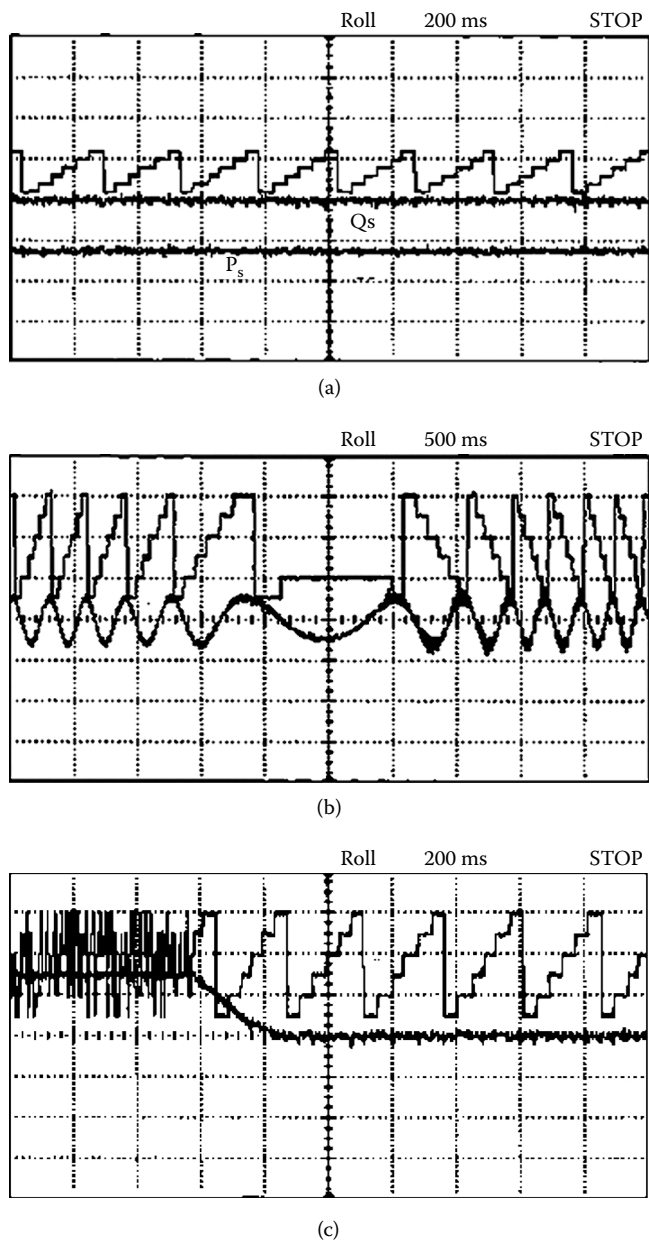


FIGURE 2.29 (a) Test results P_s , Q_s and sector step change, (b) rotor current I_r during transition through synchronous speed, and (c) Q_s slow ramping during start-up.

2.9 Independent Vector Control of Positive and Negative Sequence Currents

Unbalanced voltage conditions in the power network may occur, and they greatly influence the standard vector control of DPC schemes. It is, however, possible to superimpose vector control of direct and inverse sequence stator current, in order to gain immunity to grid disturbances. Alternatively, separate phase stator current control is feasible with one stator current as zero. A typical such scheme is presented in [Figure 2.30](#) [17].

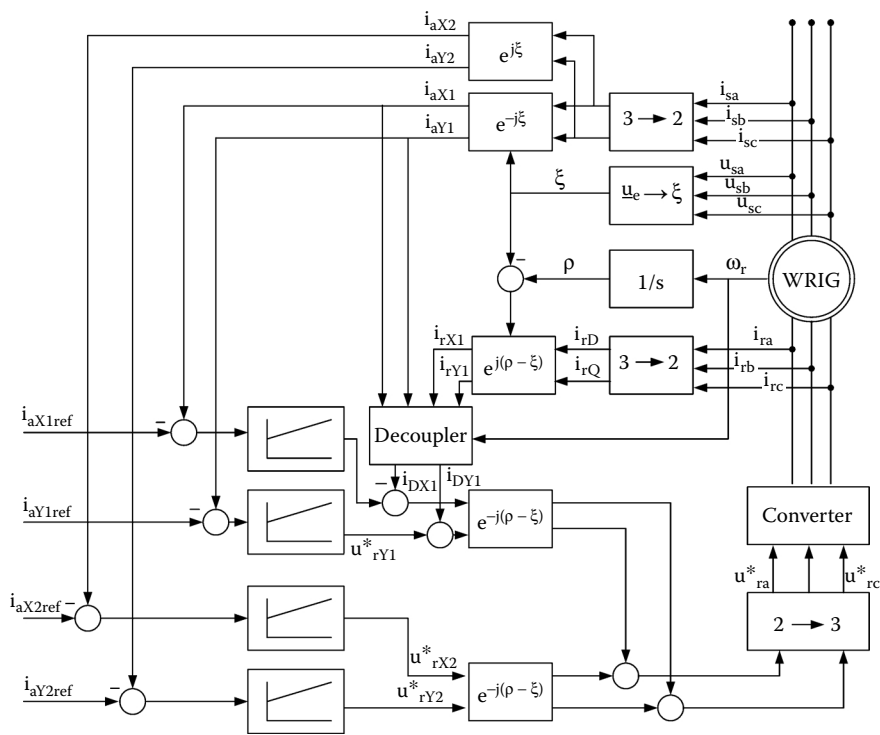


FIGURE 2.30 Positive and negative sequence current vector control of wound rotor induction generator (WRIG).

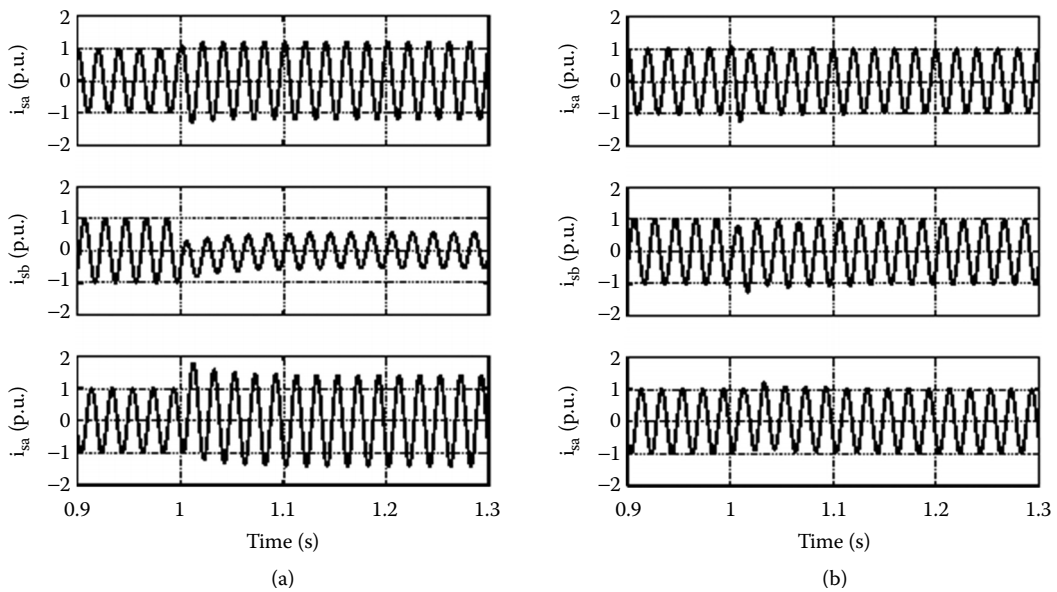


FIGURE 2.31 Response of stator phase currents with 10% negative sequence voltage: (a) positive sequence vector control and (b) positive and negative sequence control.

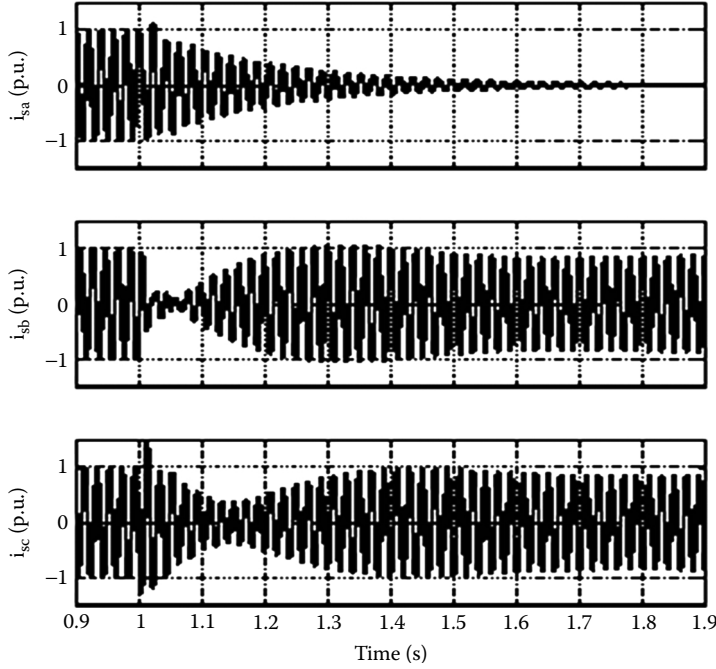


FIGURE 2.32 Stator currents in two-phase operation.

The system response in conventional (positive sequence) and, respectively, positive plus negative sequence vector control with a 10% negative-sequence voltage is shown in Figure 2.31a and Figure 2.31b, respectively, for a 200 kVA prototype [17] with $l_m = 2.9$ P.U., $l_s = l_r = 3.0$ P.U., and $r_s = r_r = 0.02$ P.U.

The reduction in stator phase current imbalance is notable. It is feasible even to drive the current in phase a to zero and work with only two active phases. As expected, the stator and rotor instantaneous active powers will pulsate, but the system is capable of feeding two-phase loads with zero current summation [17] (Figure 2.32).

These enhanced possibilities of vector control add to the power quality delivered by WRIGs.

2.10 Motion-Sensorless Control

While apparently direct power control works without a rotor position sensor, vector control requires one for Park transformations. Both schemes need speed feedback to control the prime mover. So, estimators or observers for rotor electrical position θ_{er} and speed ω_r are required for sensorless control.

The availability of stator and rotor currents through measurements is a great asset of a WRIG, for rotor position estimation. Also, remember that vector control of the machine (rotor)-side converter is performed in stator flux coordinates with

$$\bar{\Psi}_s = \Psi_s = L_s I_{ms}; \quad L_s = L_{sl} + L_m \quad (2.73)$$

where I_{ms} is the stator flux magnetization current.

In stator coordinates,

$$\bar{\Psi}_s = L_s \bar{I}_{ms} = L_s \bar{I}_s + L_m \bar{I}_r^s \quad (2.74)$$

The location of rotor current vector with respect to stator flux and phase a and rotor phase axis is shown in Figure 2.33.

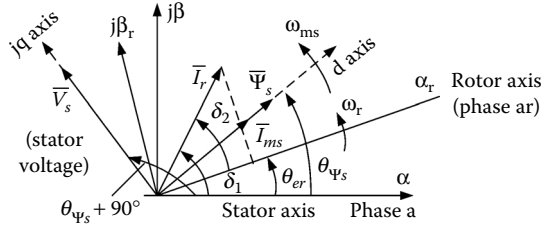


FIGURE 2.33 Location of rotor current vector with respect to stator, rotor, and stator flux axes.

The stator flux magnetization current vector $\bar{I}_{ms}$ in stator (α, β) coordinates is simply as follows (from Figure 2.33):

$$\begin{aligned} I_{ms\alpha} &= I_{ms} \cos \theta_{\psi_s} \\ I_{ms\beta} &= I_{ms} \sin \theta_{\psi_s} \end{aligned} \quad (2.75)$$

Consequently, from Equation 2.75 — in stator coordinates — the rotor current components are as follows:

$$I_{r\alpha} = (I_{ms\alpha} - I_{s\alpha}) \frac{L_s}{L_m}; \quad I_{s\alpha} = I_a \quad (2.76)$$

$$I_{r\beta} = (I_{ms\beta} - I_{s\beta}) \frac{L_s}{L_m}; \quad I_{s\beta} = \frac{1}{\sqrt{3}} (2I_b + I_a) \quad (2.77)$$

$$I_r = \sqrt{I_{r\alpha}^2 + I_{r\beta}^2} \quad (2.78)$$

But, the rotor currents are directly measured in the rotor coordinates:

$$\begin{aligned} \cos \delta_2 &= \frac{I_{r\alpha r}}{I_r} \\ \sin \delta_2 &= \frac{I_{r\beta r}}{I_r} \end{aligned} \quad (2.79)$$

with $I_{r\alpha r}$, $I_{r\beta r}$ calculated from measured i_{ar} , i_{br} , and i_{cr} :

$$I_{r\alpha r} = I_{ar}, \quad I_{r\beta r} = (2I_{br} + I_{ar}) / \sqrt{3} \quad (2.80)$$

The unknowns here are the $\sin \theta_{er}$ and $\cos \theta_{er}$, that is, the sine and cosine of the rotor electrical angle θ_{er} :

$$\theta_{er} = \delta_1 - \delta_2 \quad (2.81)$$

So,

$$\begin{aligned} \sin \theta_{er} &= \sin(\delta_1 - \delta_2) = \sin \delta_1 \cos \delta_2 - \sin \delta_2 \cos \delta_1 = \frac{(I_{r\beta} \cdot I_{r\alpha r} - I_{r\alpha} \cdot I_{r\beta r})}{I_r^2} \\ \cos \theta_{er} &= \cos(\delta_1 - \delta_2) = \cos \delta_1 \cos \delta_2 + \sin \delta_2 \sin \delta_1 = \frac{(I_{r\alpha} \cdot I_{r\alpha r} - I_{r\beta} \cdot I_{r\beta r})}{I_r^2} \end{aligned} \quad (2.82)$$

In fact, provided the stator flux magnetization current $\bar{I}_{ms}$ is known, $\sin\theta_{er}$ and $\cos\theta_{er}$ are obtained in the next time sampling step without delay. The fastest response in rotor position information is thus obtained. The rotor speed ω_r may be calculated from $\sin\theta_{er}$ and $\cos\theta_{er}$ as follows:

$$\frac{d\theta_{er}}{dt} = \hat{\omega}_r = -\sin\theta_{er} \frac{d(\cos\theta_{er})}{dt} + \cos\theta_{er} \frac{d(\sin\theta_{er})}{dt} \quad (2.83)$$

A digital filter has to be used to reduce noise in Equation 2.83 and thus obtain a usable speed signal.

Though at constant stator voltage and frequency, Ψ_s is constant, and so is I_{ms} , under faults or voltage disturbances in the power grid, Ψ_s (and thus I_{ms}) varies. It is, however, practical to start with the rated value of I_{ms} and calculate the first pair of rotor current components in stator coordinates $I'_{r\alpha}, I'_{r\beta}$:

$$\begin{aligned} I'_{r\alpha}(k) &= I_{r\alpha r}(k) \cos\theta_{er}(k-1) - I_{r\beta r}(k) \sin\theta_{er}(k-1) \\ I'_{r\beta}(k) &= I_{r\beta r}(k) \cos\theta_{er}(k-1) + I_{r\alpha r}(k) \sin\theta_{er}(k-1) \\ I'_{ms\alpha}(k) &= I_{s\alpha}(k) + \frac{L_m}{L_s} I'_{r\alpha}(k) \\ I'_{ms\beta}(k) &= I_{s\beta}(k) + \frac{L_m}{L_s} I'_{r\beta}(k) \end{aligned} \quad (2.84)$$

These new values of $(I'_{ms\alpha}, I'_{ms\beta})$ will then be used in the next time step computation cycle. So, I_{ms} is always one time step behind, but, as the sampling time is small, it produces very small errors, even if a 1 msec delay low-pass filter on $I'_{ms}(k)$ is used. The magnetic saturation curve is required to account for saturation with rotor position estimation to yield good results when the stator flux varies notably, as in grid faults. Typical rotor position estimation results are shown in Figure 2.34 [18].

The speed estimation before and after filtering is shown in Figure 2.35a and Figure 2.35b [18].

As $L_m/L_s = 1 - L_{sl}/L_s$ is close to unity, even a $\pm 50\%$ error in L_{sl}/L_s leads to negligible magnetization current estimation errors. The present θ_{er} and ω_r estimation method, with some small changes, also works when the stator currents are zero. Such a situation occurs when the synchronization conditions to connect the WRIG to the power grid are prepared. Other θ_{er} and ω_r estimators were proposed and shown to give satisfactory results [19, 20].

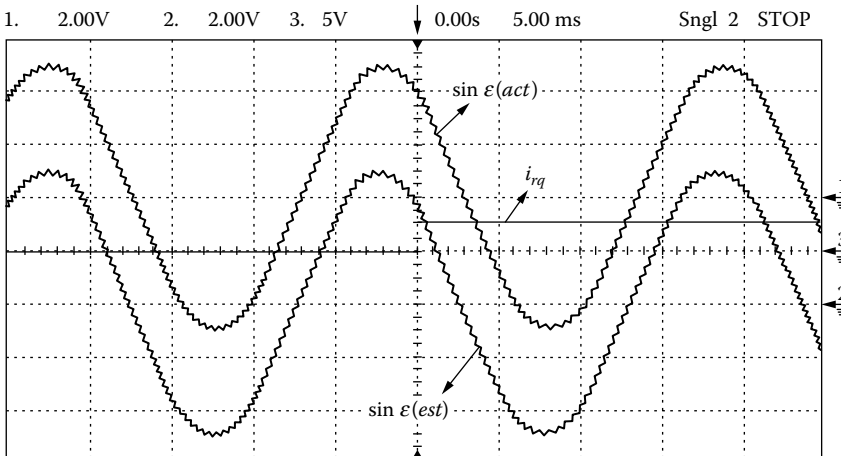


FIGURE 2.34 Estimated and actual rotor position ($\sin\theta_{er}$) for step increase V_{rq}^* .

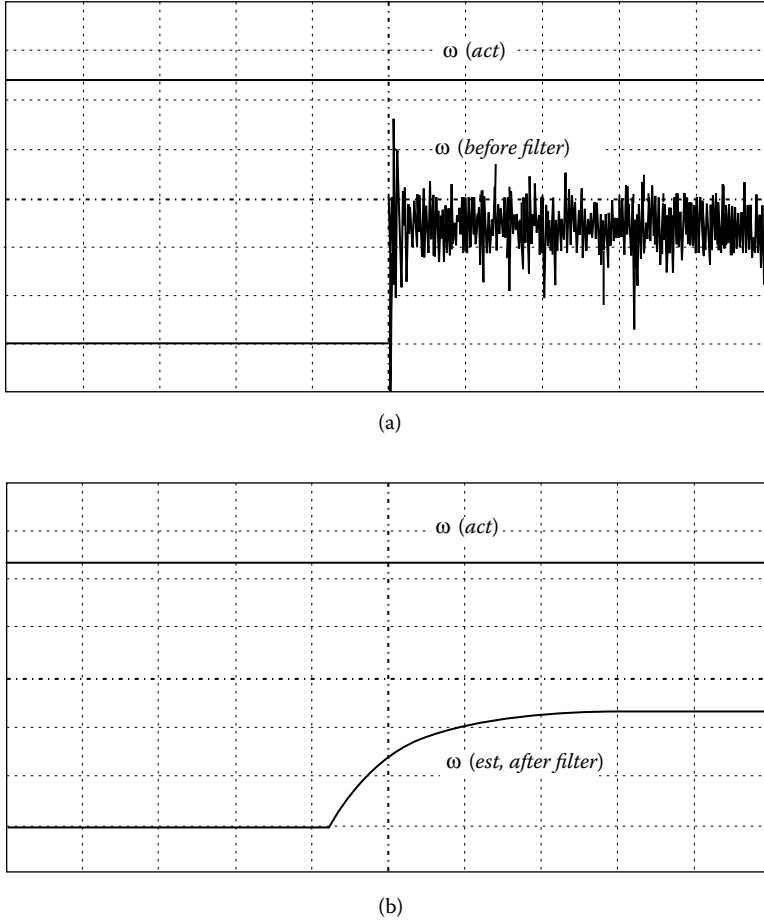


FIGURE 2.35 Estimated speed transients: (a) before filtering and (b) after filtering.

2.11 Vector Control in Stand-Alone Operation

The stand-alone operation mode may be accidental, when the power grid is not a good recipient of power, or intentional. Ballast loads are supplied in the first case. In principle, it is possible to maintain P_s , Q_s vector control as described earlier. But, the stator flux Ψ_s and frequency ω_1 are imposed, rather than estimated.

When the WRIG is isolated, self-excitation is required first, on no-load. But, after that, the vector control of the rotor source-side converter in stator voltage (or flux) orientation has a problem, as the stator voltage has harmonics even if a filter is used. Consequently, to “clean up” the noise, the stator voltage vector position θ_{V_s} is calculated as follows:

$$\theta_{V_s} = \int \omega_1^* dt + \frac{\pi}{2} = \theta_{\Psi_s} + \frac{\pi}{2} \quad (2.85)$$

where ω_1^* is the reference stator frequency. A controlled ballast load is needed to compensate for the difference between mechanical power and electrical load. The control of the source-side converter remains as that for grid operation, with the q axis (reactive power) current reference as zero, in general. The

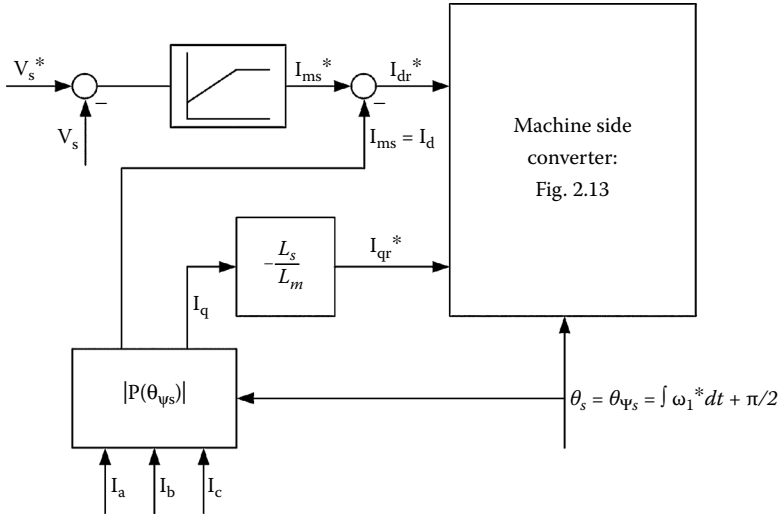


FIGURE 2.36 Machine-side converter vector control for stand-alone operation.

machine-side converter control will have the active and reactive power regulators replaced by magnetization current regulators (to keep stator voltage constant), while along axis q ,

$$I_{qr}^* = -\frac{L_s I_q}{L_m} \quad (2.86)$$

To provide for alignment along the stator flux axis,

$$\Psi_{ds} = \Psi_{\alpha s} \cos \theta_{\Psi_s} + \Psi_{\beta s} \sin \theta_{\Psi_s} \quad (2.87)$$

$$\Psi_{\alpha s, \beta s} = -\int (V_{s\alpha, \beta} - R_s I_{s\alpha, \beta}) dt \quad (2.88)$$

$$I_{ms} = \frac{\Psi_s}{L_s} \quad (2.89)$$

It would be adequate to add one more regulator for magnetization current.

The corroboration of V_s^* in the DC link with V_s^* is a matter of optimization (Figure 2.36). The turbine is commanded in the torque mode, and the torque regulator produces the control variable for ballast load to match the mechanical and electrical power when the actual load varies. More on stand-alone operation is presented in the literature [21, 22].

2.12 Self-Starting, Synchronization, and Loading at the Power Grid

There are situations, such as with pump-storage hydropower plants, when WRIG needs to start as an induction motor with short-circuited stator until slowly the speed reaches $\omega_o > \omega_{r\min} = \frac{1-S_{\max}}{\omega_1(1-S_{\max})}$. Then, the stator terminals are opened, and the machine floats with the speed slightly decreasing. During a

short period (40 to 50 msec), the synchronization conditions are prepared by adequate control. In essence, the stator voltage amplitude and phase angle differences (errors) are driven to zero by close-loop control of I_{dr} and I_{qr} — the rotor current components. In the process, the rotor phase sequence used for starting to ω_o has to be reserved to prepare for synchronization after the stator windings are opened.

So, instead of active and reactive power regulators (Figure 2.13), the voltage and phase angle error regulators are introduced (Figure 2.37a).

Fast synchronization is noticed in Figure 2.37b[19]. Starting as a motor with the short-circuited stator may be done with or without an autotransformer. The autotransformer is disconnected at a certain speed ω_{ms} above a certain synchronization speed ω_o , in order to take advantage of higher voltage and, thus, allow for a lower rating static power converter to be connected to the rotor (Figure 2.38a and Figure 2.38b) [19].

Control during motor starting with a short-circuited stator may be done, and also with the vector control scheme of Figure 2.21, only it must be slightly modified. For example, the reactive power regulator will be bypassed. The reactive rotor reference current I_{dr}^* is kept constant and positive as the machine magnetization is produced through the rotor. The current I_{dr}^* should correspond to the rated no-load current of the WRIG. The active power regulator in Figure 2.13 should be replaced by a speed regulator (Figure 2.39).

To illustrate the transient response of WRIG for generating and pumping operation at the power grid, results for a 400 MW unit are shown in Figure 2.40 and Figure 2.41 [22]. This large WRIG was also started with a shorted stator for the pumping operation mode.

The response is smooth and fast, proving the vector control capability to handle decoupled active and reactive power control.

2.13 Voltage and Current Low-Frequency Harmonics of WRIG

Current harmonics in a WRIG originate mainly from the following:

- Grid voltage harmonics
- Winding space harmonics
- Switching behavior of the static power converter connected to the rotor

The high-frequency harmonics are left aside here. As the power grid is connected both to the stator (directly) and to the rotor (through the AC–AC converter), the voltage harmonics influence the stator and rotor currents. The converter has to handle fundamental power control and also act as a filter for these harmonics.

Alternatively, active filters may be placed between the power grid and the nonlinear loads [23]. In what follows, the low-frequency source current and equations are provided and used as a basis for feed-forward control added to fundamental vector control of a WRIG to attenuate current harmonics.

The significant voltage harmonics due to the power grid are of the order $6k + 1$ (positive sequence) and $6k - 1$ (negative sequence) in stator coordinates and $\pm 6k$ in stator voltage fundamental vector coordinates:

$$V_s = \hat{V}_{s1} \left(1 + \sum_k C_{-6k-1} e^{-6jk\omega_1 t} + \sum_k C_{-6k+1} e^{+6jk\omega_1 t} \right) \quad (2.90)$$

The distorted supply voltage produces current harmonics through the main (sine filter) choke (L_c , R_c) placed in series on the source-side (front end) converter. They are driven by the voltage drop over the main choke:

$$V_{sv} - V_{invv} = R_c I_{s,v} + j\omega_{sv} L_c I_{s,v} \quad (2.91)$$

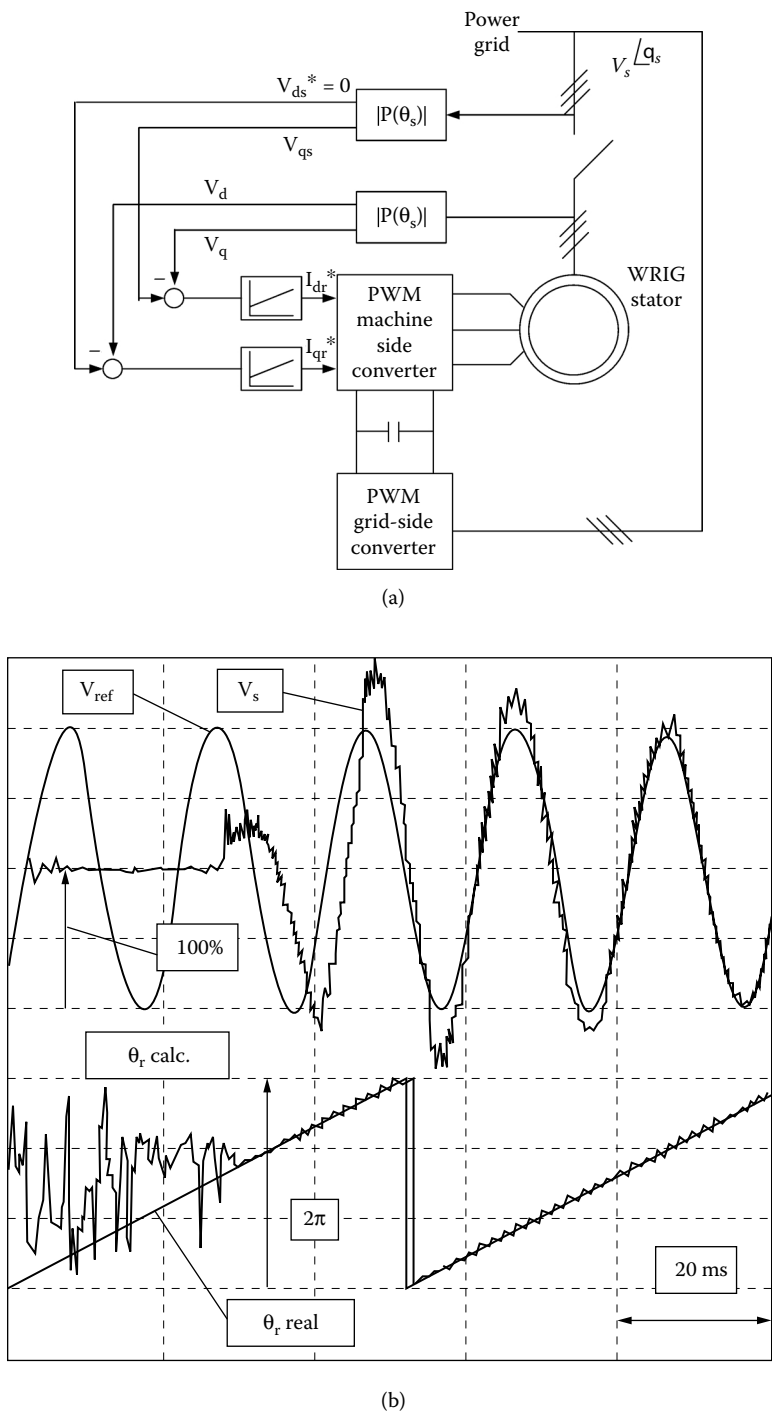


FIGURE 2.37 Mode II (synchronization): (a) control scheme and (b) results.

where V_{sv} , V_{invv} are, respectively, the vectors of harmonics in supply-side and source-side converter output voltage. On the other hand, the nonsinusoidal distribution of WRIG windings (placed in slots) produces rotor flux harmonics of $6k + 1$ orders. In rotor coordinates, they are of the following form:

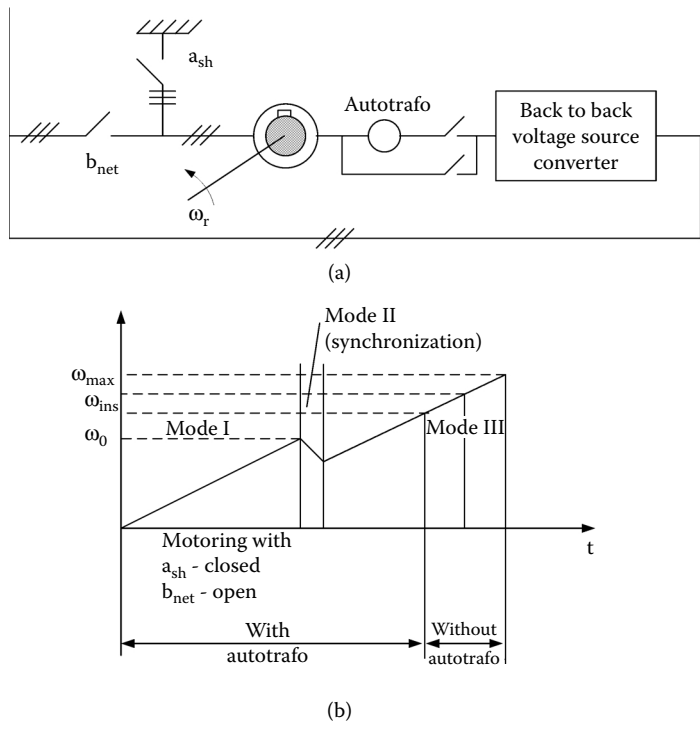


FIGURE 2.38 The three operation modes (a) and the speed vs. time (b).

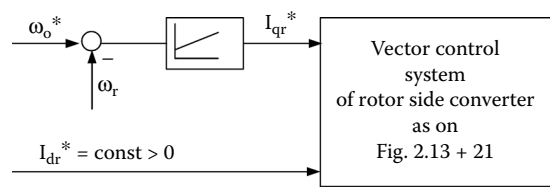


FIGURE 2.39 Rotor reference currents during shorted-stator motor starting to speed $\omega_o > \omega_{rmin}$.

$$\Psi_r^r = \Psi_{r1} \left(1 + \sum_k C_{\Psi_{6k-1}} e^{-6jk\omega_r t} + \sum_k C_{\Psi_{6k+1}} e^{+6jk\omega_r t} \right) e^{j(\omega_1 - \omega_r)t} \quad (2.92)$$

They produce stator current harmonics of the same order. In stator coordinates,

$$\bar{I}_s^s = I_{s1} \left(1 + \left(\sum_k C_{i_{6k-1}} e^{-6jk\omega_r t} + \sum_k C_{i_{6k+1}} e^{+6jk\omega_r t} \right) e^{j(\omega_1 - \omega_r)t} \right) \quad (2.93)$$

The amplitude of stator current harmonics thus produced depends heavily on the machine winding type and on power level. They tend to decrease in large power machines but are still non-negligible.

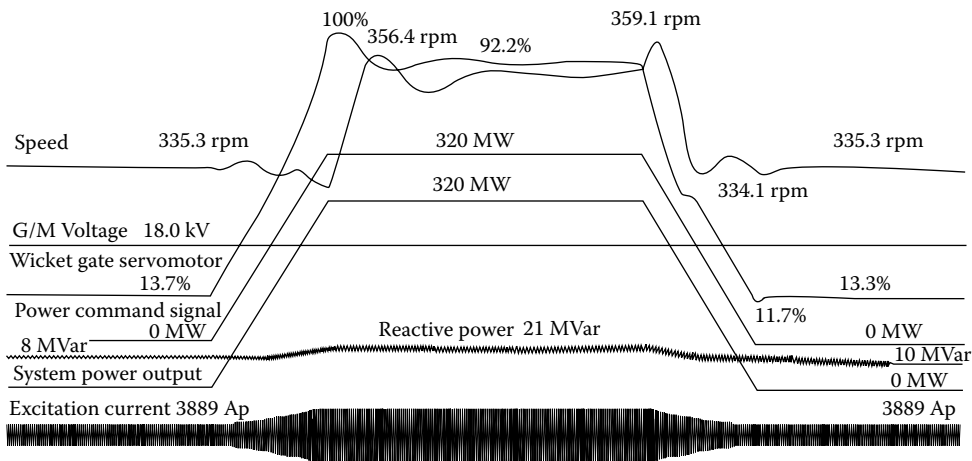


FIGURE 2.40 Wound rotor induction generator (WRIG) (Ohkawachi unit 4) ramp response for generating mode.

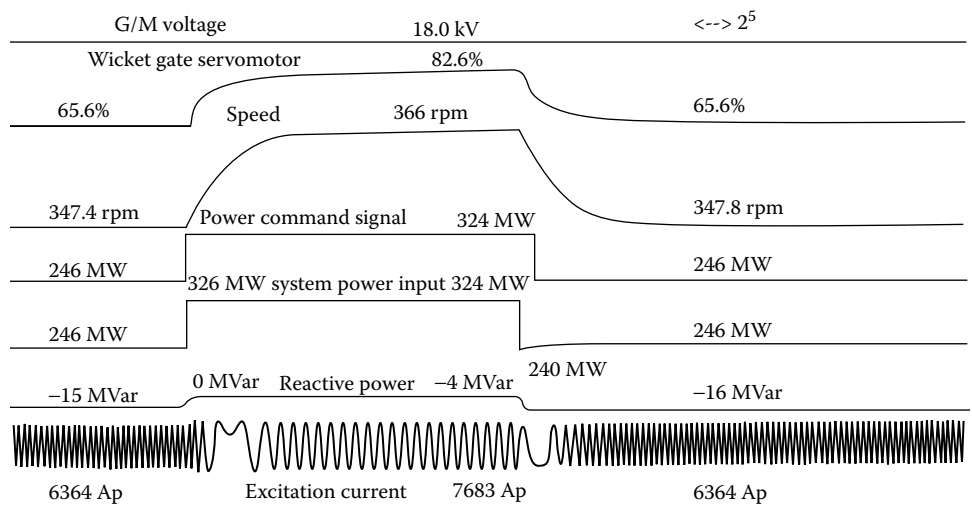


FIGURE 2.41 Wound rotor induction generator (WRIG) (Ohkawachi unit 4) ramp response for pumping (motoring) mode.

In order to compensate these harmonics of stator current, the machine behavior toward them should be explored. Rewriting the space-phasor equation for the fundamental (Equation 2.15) in stator coordinates ($\omega_b = 0$), for rotor flux Ψ_r and stator current I_s vectors as variables, we obtain the following:

$$\bar{V}_s - \bar{V}_r \frac{L_m}{L_r} = \left(R_s + \frac{L_m^2}{L_r^2} R_r \right) \bar{I}_s + L_{sc} \frac{d\bar{I}_s}{dt} + \frac{L_m}{L_r} \left(j\omega_r - \frac{R_r}{L_r} \right) \bar{\Psi}_r \quad (2.94)$$

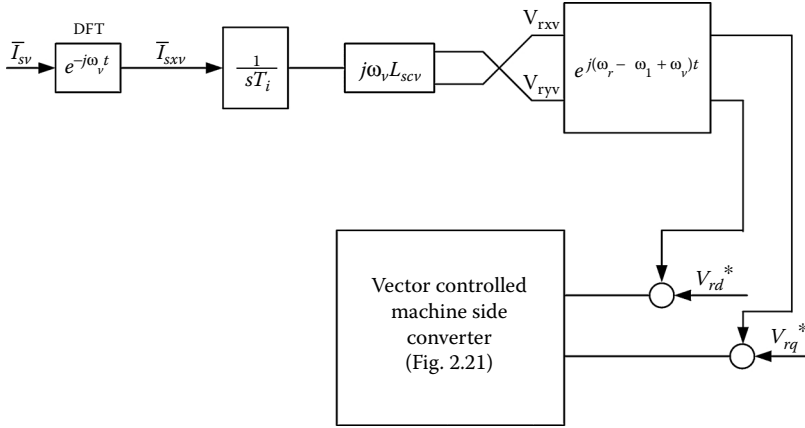


FIGURE 2.42 Feed-forward compensation stator current harmonic v .

The same equation may be applied for current (voltage) harmonics, provided the influence of frequency on resistances and inductances is considered (R_v, L_{scv}). As the rotor flux varies slowly (with the time constant $T_r = L_r/R_r$), the stator voltage harmonics are not followed by $\hat{\Psi}_r$, which then may be eliminated from Equation 2.94:

$$\bar{V}_{s,v} - \frac{L_m}{L_v} \bar{V}_{r,v} \approx (R_v + jv\omega_1 L_{scv}) \bar{I}_{s,v} \quad (2.95)$$

On the other hand, harmonics originating from rotor flux are not reflected in the stator voltage, and Equation 2.94 gets the following form:

$$\frac{L_m}{L_r} \left(-j\omega_r + \frac{R_r}{L_s} \right) \bar{\Psi}_{r,v} - \frac{L_m}{L_r} \bar{V}_{r,v} = (R_{vr} + jv\omega_r L_{scvr}) \bar{I}_{s,v} \quad (2.96)$$

These rotor-flux-produced stator current harmonics only slightly depend on speed, as speed occurs on both sides of Equation 2.96. However, V_r increases with speed almost linearly in WRIG. Consequently, the compensation voltage, equal and opposite to the left side of Equation 2.96, is proportional to speed. The elimination of stator current harmonics in Equation 2.95 and Equation 2.96 should be done by making the voltage difference on their left side zero by proper compensation. Feed-forward compensation is a favored such method. Advantages of Equation 2.91, Equation 2.95, and Equation 2.96 are taken in such a typical scheme, required for each harmonic (Figure 2.42).

The output of the feed-forward controller enters the vector control rotor (machine)-side converter at the rotor voltage references level in synchronous coordinates. It goes without saying that the controller bandwidth has to be increased to cope with these new higher frequency (hundreds of hertz) components. Typical results with 70 to 80% compensation of fifth and seventh harmonics are shown in Figure 2.43a for a 4 kW WRIG and in Figure 2.43b for a 2.5 MW WRIG [24]. Though the fifth and seventh current harmonics are larger for the 4 kW machine, their reduction is also worthwhile in the 2.5 MW machine to improve power quality.

Note that converter-produced harmonics are in the hundreds of hertz for thyristor rectifier-current source inverter converter WRIGs and in the kilohertz range for DC voltage link PWM AC-AC converters with IGBTs or IGCTs. They strongly depend on the control and converter switching (PWM) method. And, they are larger and more damaging in DC current link AC-AC converter WRIGs [25]. In DC voltage link PWM AC-AC converters, the main harmonics are around the fixed (if fixed) pulse frequency and

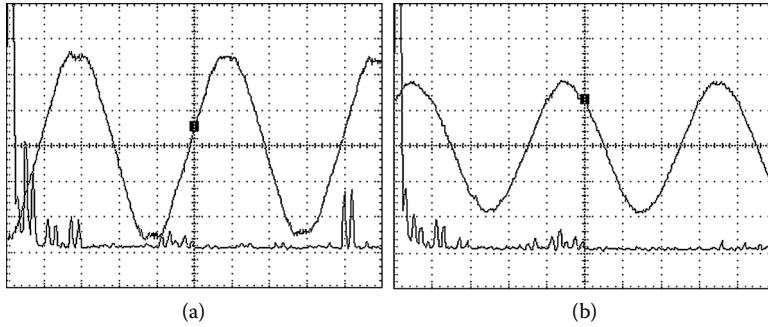


FIGURE 2.43 Stator current harmonic feed-forward compensation: (a) 4 kW wound rotor induction generator (WRIG) and (b) 2.5 MW WRIG.

its multipliers. Multiple-level voltage PWM AC–AC indirect converters are a good solution for raising the pulse frequency, as seen in the rotor currents.

But essentially, the reduction of converter produced harmonics takes place with the use of the PWM techniques used for the scope. Random switching or randomized switching frequency or pulse positioning are favored methods to reduce the converter-caused line current harmonics. A thorough analysis of various PWM techniques with their effects on stator and rotor current harmonics is given in Reference [26]. The common modulation plus triple sine is found to be appropriate. Also, shifting the pulse period of the rotor-side PWM converter by a quarter of a period with respect to the source-side PWM converter produces a reduction of the harmonics spectrum at the double pulse frequency. Passive or active filters may also be used for the scope.

2.14 Summary

- WRIGs are also called DFIGs or DOIGs or even doubly fed SGs.
- The uniformly slotted cylindrical laminated stator and rotor cores of WRIG are provided with three-phase AC distributed windings.
- For steady state, the mmfs of the stator and rotor currents both travel at the electrical speed of ω_1 , with respect to the stator. However, with electrical rotor speed ω_r , the frequency ω_2 of rotor currents is $\omega_2 = \omega_1 - \omega_r$. Negative ω_2 means negative slip ($S = \omega_2/\omega_1$) and an inverse order of phases on the rotor.
- With a proper power supply connected to the rotor at frequency ω_2 (variable with speed such that $\omega_1 = \omega_r + \omega_2 \approx \text{constant}$), the WRIG may work as a motor and generator subsynchronously ($\omega_r < \omega_1$; $\omega_2 > 0$), supersynchronously ($\omega_r > \omega_1$; $\omega_2 < 0$), and even at synchronism ($\omega_r = \omega_1$, $\omega_2 = 0$). Constant stator voltage and frequency may be secured for variable speed: $\omega_1(1 - |S_{\max}|) < \omega_r < \omega_1(1 + |S_{\max}|)$. The larger the speed range, the larger the power rating P_{rN} of the rotor-connected static power converter: $P_{rN} \approx |S_{\max}| P_{sN}$.
- For $\pm 25\%$ slip and speed range, the power rating of the rotor-connected static power converter is around $25\% P_{sN}$, that is, 25% of stator power. The WRIG may deliver 125% total power at 125% speed, with a stator designed at 100% and 100% speed. The flexibility of a WRIG due to variable speed and the reasonable costs of the converter are the main assets of WRIGs.
- The phase-coordinate model of the WRIG has an eighth order, and some coefficients are rotor-position (time) dependent. It is to be used in special cases only.
- The space-phasor or complex-variable model is particularly suitable for investigating WRIG transients and control. Decomposition along orthogonal axes, spinning at general speed ω_b , leads to the d – q model. The Park generalized transform relates the phase coordinate to the space-phasor model. The latter has all coefficients independent of rotor position.

- The active power is positive (delivered) in the stator for generating and negative for motoring, for both $S > 0$ and $S < 0$.
- The active power is positive (delivered) in the rotor for motoring (for $S > 0$) and negative (absorbed) for generating. For $S < 0$, the reverse is true: the motor absorbs power through the stator and the rotor, and the generator delivers it through the stator and the rotor.
- With losses neglected, the mechanical power $P_m = P_s + P_r$.
- The space-phasor model of the WRIG, for steady state, in synchronous coordinates, is characterized by DC quantities (voltages, currents, and flux linkages) that make it suitable for control design.
- The space-phasor model is characterized by its equations, space-phasor diagrams, and structural diagrams.
- WRIG transients are to be approached via the space-phasor model. For scalar open-loop control at constant rotor voltage V_r or current I_r , the linearization of the d - q model leads to a sixth-order equation to determine the complex eigenvalues. The order is reduced to four if the stator resistance R_s is neglected.
- It has, by now, been shown that only rotor current control provides for a stable motor and generator, both undersynchronously ($S > 0$) and supersynchronously ($S < 0$).
- As WRIGs are supplied, in general, in the rotor, by static power converters, and vector or direct power or feedback linearized control is applied, the investigation of transients of the controlled WRIG becomes most relevant.
- Static power converters for the rotor circuit of a WRIG may be classified as follows:
 - DC current link AC-AC converters
 - DC voltage link AC-AC converters
 - Direct AC-AC converters (cycloconverters and matrix converters)
 - They all may be built to provide bidirectional power control, both for $S > 0$ and $S < 0$. However, the DC current link AC-AC converter fails to work properly very close to or at synchronism ($\omega_r = \omega_1$). The content in current harmonics depends both on the converter type and on its PWM and control strategies. The DC voltage link AC-AC (back-to-back) converter with IGBTs, GTOs, or IGCTs seems to be the way of the future.
- Vector control of a WRIG refers, separately, to the machine-side converter and to the supply-side converter.
- The vector control of a machine-side converter essentially uses active and reactive power closed-loop regulators to set reference rotor d - q current components i_{qr} and i_{dr} in stator flux synchronous coordinates ($\omega_b = \omega_1$). After voltage decoupling, the rotor voltage components V_{dr}^* , V_{qr}^* are Park-transformed into rotor coordinates, to produce the reference rotor voltages V_{ar}^* , V_{br}^* , and V_{cr}^* . A PWM strategy is used to “copy” these patterns.
- Vector control of the source-side converter also works in synchronous coordinates but only if aligned to stator voltage (90° away from the stator flux axis). In essence, the d axis source-side current i_d is used to control the DC link voltage (active power) through two closed-loop cascaded regulators. The q axis source-side current i_q is set to provide a certain power factor angle on the source side, through a current regulator. The i_{dr} , i_{qr} components are then Park-transformed in stator coordinates to produce the source-side voltages by the source-side converter.
- Vector control was successfully used for fast active and reactive power control at the power grid and in stand-alone operation.
- Even after a three-phase short-circuit, the WRIG recovers swiftly and with small transients, if the reference rotor currents on the machine-side converter are limited by design.
- Besides vector control, for the machine-side converter, direct power control (DPC) was proposed. DPC stems from direct torque control applied to AC drives.
- DPC uses, in principle, hysteresis active and reactive power regulators to directly trigger one (or a sequence of) voltage vector(s) in the machine-side converter. A kind of random PWM is obtained. DPC claims simplicity for fast and robust response in power and implicit motion-sensorless control.

- Vector control requires rotor electrical position θ_{er} and speed ω_r information for implementation. In addition to using sensors to cut costs, motion sensorless control is used.
- The θ_r and ω_r observers are easier to build for a WRIG, as both stator and rotor currents are measurable, though, each in its coordinates. However, by estimating the stator flux, it is easy to use the current model and estimate $\cos\theta_{er}$ and $\sin\theta_{er}$, and then ω_r . Low-pass filtering is mandatory for both rotor position and rotor speed estimation. Good sensorless operation during synchronization to, and operation at, the power grid was demonstrated.
- There are applications where self-starting (and motoring, for pump storage) is required. It is done with the stator short-circuited, while the active power regulator is replaced by a speed regulator, and the reactive power regulator is replaced by constant reference i_{dr}^* current.
- During the synchronization mode, the errors between d - q WRIG and power supply voltages are driven to zero by closed-loop regulators that replace the P_r and Q_r regulators. All of these are in the machine-side converter.
- Vector control may be extended to include the negative sequence components and may, thus, enable the handling of asymmetrical power grids, up to zeroing a one-phase current.
- In stand-alone operation, the vector control also works, but the source-side converter has terminal voltage control at a given frequency. The DC link voltage will float with the terminal voltage. The presence of the filter somewhat complicates the control. Smooth passage, from grid to stand-alone operation, is feasible.
- Smooth and fast stator active and reactive power control with WRIGs was demonstrated up to 400 MW/unit.
- The static power converter, the distribution of coils in slots in the WRIGs, and the power grid voltage cause (or contain) harmonics.
- The switching harmonics due to PWM in the converter may be attenuated by adequate, quarter of a period delays of pulses on the two sides of the back-to-back converters or through special active filters or by using random PWM.
- The current harmonics due to stator voltage and rotor distributed windings may be compensated one by one by adding their compensating voltages to V_{ds}^* and V_{qr}^* in the standard machine-side converter vector control scheme. Alternatively, active filters may be used for the scope.
- The power quality of a WRIG can be made really high, and efforts and solutions in that direction are mounting.
- Worldwide research efforts are dedicated to investigating stability in power systems with WRIGs driven by wind or hydraulic turbines [2].

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